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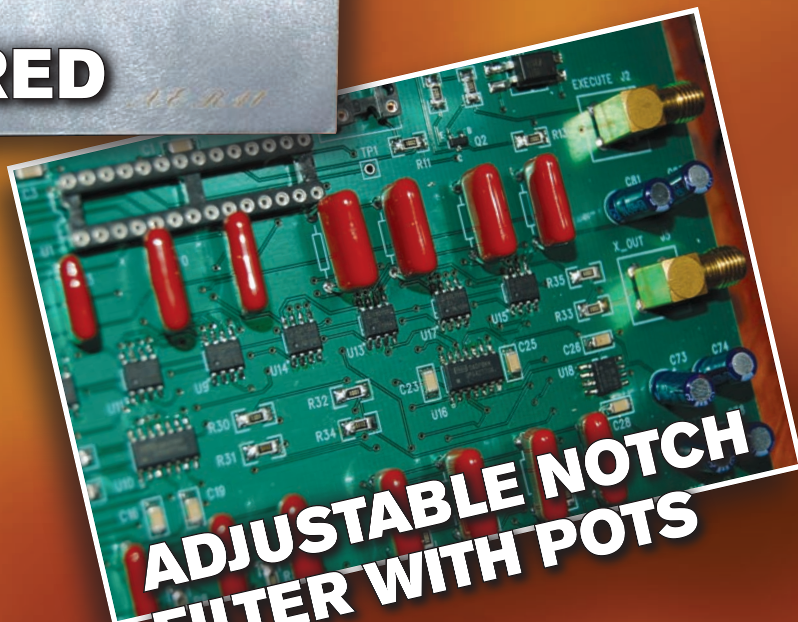

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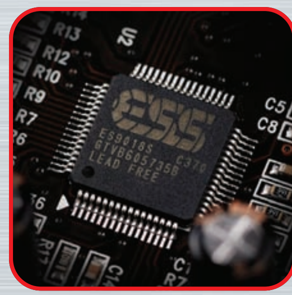

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Farewell

Dear Friends,

Audio Amateur Inc. is in a wonderful transition. By the time you read this, all of our assets will have been acquired by the Elektor group, which publishes *Circuit Cellar* and the North American version of *Elektor*. This ensures a more firm, well-funded publishing organization behind our periodical and book publishing activities. The buyers have major plans to strengthen and expand *audioXpress*, *Voice Coil*, and our two directories—the annual *Loudspeaker Industry Sourcebook* and the biennial *World Tube Directory*—as well as our list of books and CDs.

On a personal note, some of you will realize that I founded the company in 1970, and that we are now in our 41st year of operation. Birthdays pile up. Next February I will be 89. I am in fine health and still active. But it is time to hand the reins to others. I have been a publisher for a long time and my love for the profession continues undiminished. I am deeply pleased about the plans and enthusiasm Elektor is bringing to our enterprise.

One of the privileges I will miss the most is my ongoing relationship with the more than 900 authors who have made our publications possible. Of these, over a third are from countries outside of the United States. From the beginning, over 40 years ago, authors have offered us outstanding fare. Just yesterday I received a new manuscript from our star performer, Nelson Pass. I take this as a good omen that the quality of these periodicals will continue.

I am pleased to report that Vance Dickason, whose talents have been a very significant factor in our success, will continue as editor of *Voice Coil*. We are all working hard in this transition hoping that all will go smoothly. We ask for your patience during this period.

Please accept my thanks for the pleasure of being publisher for an audience who have responded in all sorts of wonderful ways over our years together.
—E.T.D.

Designing Amps for **Maximum Linearity** with the **ECC99**

The author shares his design methodology for a simple tube amp.

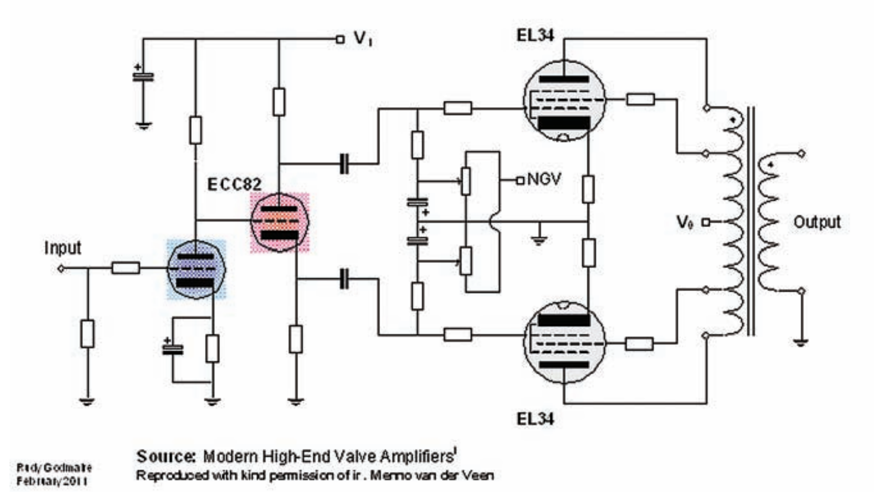

FIGURE 1: The VDV amplifier concept.

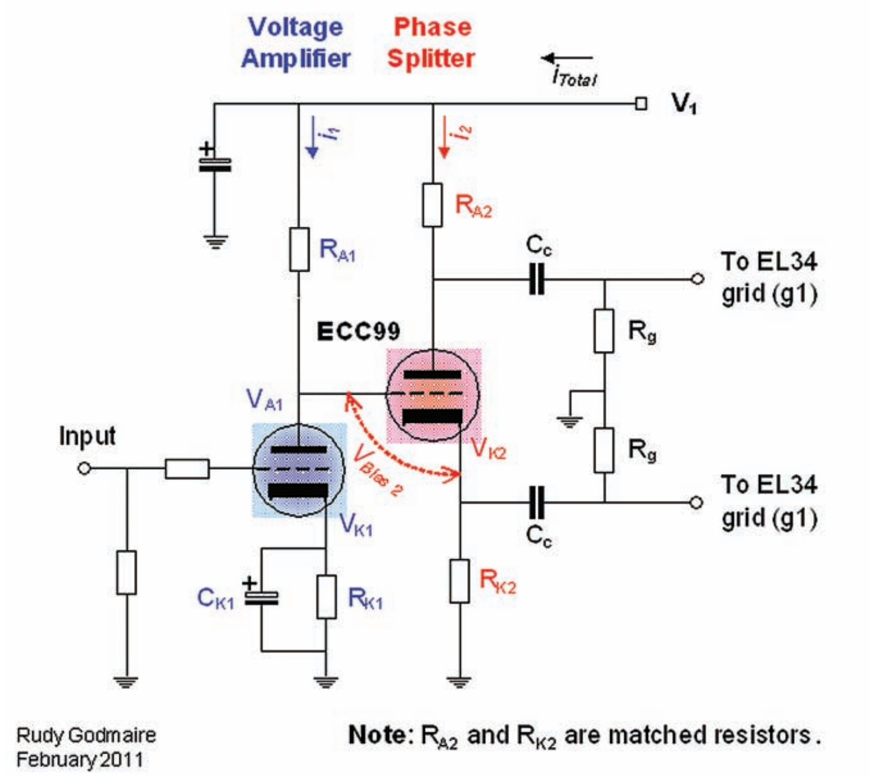

FIGURE 2: The input tube circuitry.

I always thought that designing a great tube amplifier was a mixture of science and art, and I still do. There are numerous topologies that can achieve high fidelity, some rather complex and others quite simple. The circuitry I use to drive the push-pull output stage of my VDV amplifiers certainly belongs to the latter category. **Figure 1** shows a generic view of the topology designed by Menno van der Veen and explained in his book *Modern High-End Valve Amplifiers*[1]. I like the minimalism of this circuit very much because it enables excellent musicality.

In this article, I will thoroughly explain the methodology I developed to design and optimize the input circuitry of such amplifiers using an ECC99 in place of the ECC82. I will use the case of my own VDV amplifiers as an example. I believe even experienced designers will be pleased with this protocol because it reduces the drawing process with loadlines to a maximum of three iterations. Note that the proposed methodology works also very well with other tubes having low plate resistance such as the ECC88/6DJ8 and the 6H30pi.

OVERVIEW

The input circuitry of my amplifiers is designed around the ECC99[2] manufactured by JJ Electronic in Slovakia (www.jj-electronic.com). This medium mu double triode tube offers some very nice features with its relatively high transconductance (S = 9.5mA/V) and its low plate resistance (r_p = 2300Ω). The amplification factor (μ = 22) of this tube constitutes a great advantage for the topology presented in **Fig. 1**.

Figure 2 focuses on the circuitry surrounding the input tube. The first half of

the ECC99 is a voltage amplifier stage whose function is to amplify the signal coming from the source. With its cathode bypassed by a capacitor, the gain is perfect to drive power tubes such as the EL34 and KT77. This stage is direct-coupled to the second half of the tube that acts as a split-load phase splitter (Concertina) and driver stage for the power tubes.

Direct coupling requires that the anode of the voltage amplifier must have a DC potential (V_{A1}) slightly lower than the DC potential of the cathode of the phase splitter (V_{K2}). The difference between these potentials sets the bias of the phase splitter ($V_{Bias\ 2}$). Coordinating these two voltages while maintaining the same supply voltage (V_1) for both triodes constitutes the main design challenge of this circuit.

On another front, when you look deeper into the way this circuit functions, you realize that when i_1 increases, i_2 decreases, and vice versa. As stated by Morgan Jones in his book *Valve Amplifiers*[3]: "Ideally, we would juggle with the signal current in the input stage to be equal and opposite to that in the Concertina. . ." Doing so makes i_{Total} remain constant all the time. This way of designing the circuit is easier than it appears, and if you are aiming at the very best design possible, I believe this extra care is really worth it.

Before launching into the main development of this article, I must mention that fidelity is always my main leitmotif when I design a circuit. This means I conscientiously endeavor to set the parameters in the most linear zone of the tube, usually at the expense of voltage gain. Personally, I find that high fidelity is better achieved when you favor simple circuits and use them well within their limits.

All the considerations discussed so far are addressed in the methodology I developed and that I will now expose. It is derived from the best practices I have learned while studying the works of experts such as Chaffee[4], Crowhurst[5], Langford-Smith[6], Jones, Reich[7], Rozenblit[8], and van der Veen.

THE OUTPUT STAGE

Because the following stage always determines the requirements of the preceding stage, the design of a power amplifier is realized going backward, from the output to the input. As far as the phase splitter is concerned, what interests me is the grid voltage swing that is required to drive the output tubes to full power.

The output stage of my amplifiers uses EL34 power tubes. With a plate voltage of 365V, you can normally set the quiescent current to a maximum of 68mA (25W/365V). However, because they are connected to the output transformers in the triode mode, the EL34s can be biased higher up to a maximum of 82mA. This can be explained by the fact that the maximum permissible dissipation of the plate and screen grid together ($P_a + P_{g2}$) is 30W[9]. However, biasing the tubes near their absolute maximum reduces their lifespan, which is why I prefer a more conservative value. In my VDV amplifiers, I set the bias to 70mA.

Inspection of the Mullard EL34 datasheet reveals that the grid bias voltage V_{g1} is -27V for a plate voltage of 365V and a quiescent current of 70mA[10]. Because I have a push-pull topology, the phase splitter must provide a voltage swing of at least ±27V with minimum distortion. As you will realize later, there is no need to add a safety margin now to this voltage swing evaluation because the way I design my circuits leaves plenty of headroom.

SPLIT-LOAD PHASE SPLITTER (CONCERTINA)

It is usually accepted that the higher the supply voltage of the phase splitter, the better. As you can see in **Fig. 3**, while loadlines #1 and #3 have the same load resistance, it appears obvious that loadline

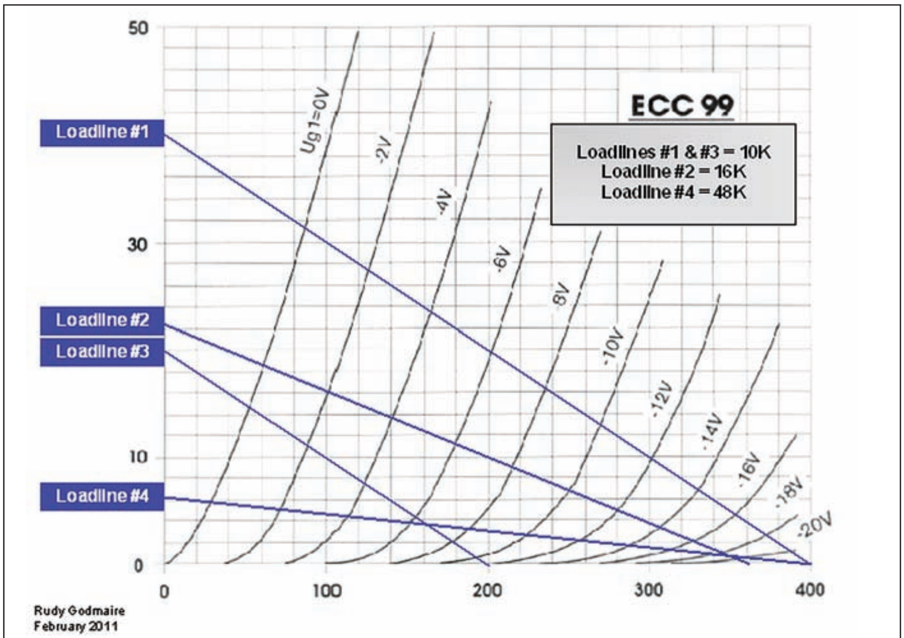

FIGURE 3: Linearity of various loadlines.

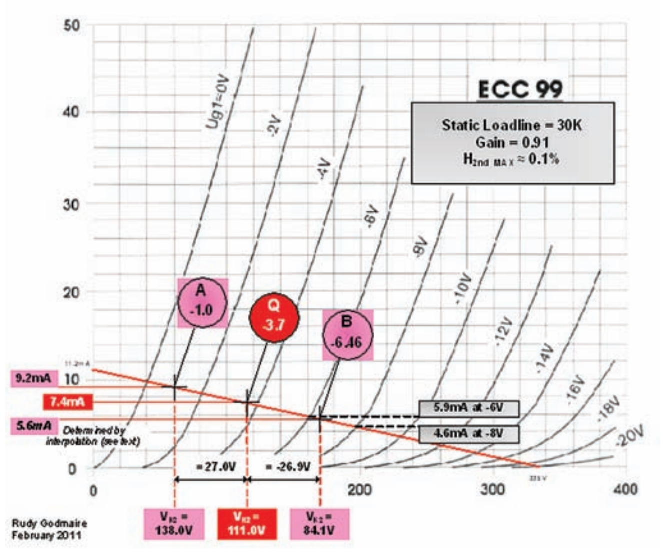

FIGURE 4: Phase splitter loadline.

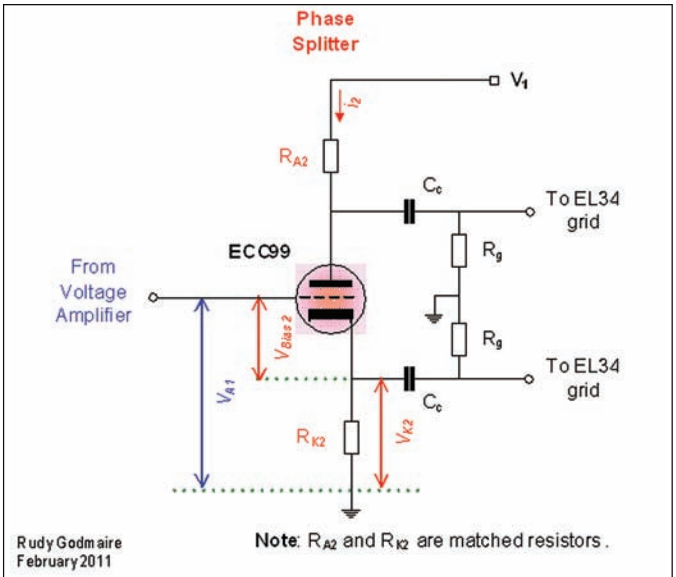

FIGURE 5: Phase splitter circuitry.

#1 provides better linearity. With a high tension of 400V and a load resistance of 10kΩ, the grid voltage lines show fairly even spacing from 0V to -10V.

On the other hand, maintaining a high supply voltage while significantly increasing the value of the load resistance flattens the loadline to a point where it compromises linearity (see loadline #4). Designing an efficient phase splitter in the most linear zone thus implies aiming at the highest voltage possible while using an optimum load resistance that will not draw unnecessary current, such as in loadline #2.

My experience with the ECC99—and also with the ECC88/6DJ8 and the 6H30pi—reveals that excellent linearity and efficiency are obtained when the value of the anode load (R_{A2}) and cathode load (R_{K2}) is set between five to ten times the value of the plate resistance r_p of the tube. Knowing that r_p is around 2300Ω for the ECC99, the value of R_{A2} and R_{K2} will range between 11.5kΩ and 23kΩ.

Designing for maximum linearity also implies that you consider another form of very objectionable distortion the graph does not reveal: positive grid current distortion. Even though the loadlines might suggest excellent linearity when the grid voltage is between 0V and -1V, you must avoid using the circuit in this area because this is where grid current starts to occur. The exact point where this begins, which is called contact potential, is not mentioned in the ECC99 datasheet nor in many other small tube datasheets. This is why it is wise to consider -1V as the maximum value of grid voltage. This determines the upper limit of the grid voltage swing of all my designs with small tubes, be it a phase splitter, a voltage amplifier, or any other topology.

Figure 4 shows the plate diagram of the phase splitter. I have set R_{A2} and R_{K2} to 15kΩ because this value provides the best optimization in my circuit. Note that it is highly recommended that these resistors be matched to ensure perfect balance of both output legs. The static loadline that appears on the graph is therefore a straight line that represents all the operating points that can be assumed with the total DC load resistance of 30kΩ and the plate supply voltage of my circuit (V_1 = 335V), when there is no signal applied to the grid of the phase splitter.

From there, I first set the -1V upper limit on the slope (point A). Projecting this point on the current axis indicates that the circuit will draw 9.2mA, which will result in a voltage drop of 138V across $R_{K2.}$

At the quiescent point (point Q), you know that the voltage across V_{K2} must decrease by 27V as compared to point A to drive the EL34s to full power. Dividing 27V by R_{K2} converts this voltage into the corresponding variation of current of 1.8mA. This sets the quiescent current to 7.4mA and the voltage drop through R_{K2} to 111.0V. Carefully projecting this value on the DC loadline also determines the bias of the phase splitter, which will be -3.7V.

Determining the lower limit of the grid voltage swing (point B) requires some calculation. If the characteristic curves were perfectly linear, point B would be at -6.4V. Unfortunately, that is not the case because as the grid voltage becomes more negative, the grid lines become closer to each other. In reality, point B will be slightly more negative and you must find its value by interpolation.

You will see that the split-load phase splitter is a special case of the cathode degenerative triode, in which the cathode and anode resistors exhibit the same value. Some methods enable the calculation of the gain and distortion of such a circuit. The following development is based on Krauss' method[11].

Figure 5 shows a different view of the phase splitter. From this schematic, it becomes quite obvious that

(1) $V_{A1} = V_{Bias\ 2} + V_{K2}$

Using a spreadsheet, I first computed the values for $V_{Bias\ 2}$, i_2, V_{K2}, V_{A1}, at points A and Q (**Table 1**). Then I calculated the absolute variation of V_{K2} and V_{A1} around the quiescent point which appears as ΔV_{K2} and ΔV_{A1} in the table. If you make the hypothesis that the input signal is undistorted, you know that ΔV_{A1} will be identical either side of point Q. Consequently, point A being at 29.7V, point B will need to be at -29.7V.

You can also deduce that point B will lie somewhere between the -6V and -8V grid voltage lines in **Fig. 4**. By comput-

ing the values of i_2 for these operating points in the spreadsheet, it becomes easy to calculate point B by interpolation. This is done by first determining an interpolation ratio this way:

ΔV_{A1} at Point B': -24.8V
ΔV_{A1} at Point B: -29.7V   4.9V
21.5V
ΔVA1 at Point B": -46.3V

Interpolation Ratio:
4.9V/21.5V = 22.8%

Once this ratio is known, you simply use it to interpolate the values of V_{Bias2} and i_2 at point B, which will allow the calculation of V_{K2}, ΔV_{K2}, and V_{A1}.

Table 1 provides all the information you need to calculate the gain of the phase splitter and also its level of second-order harmonics.[12] You find the gain by dividing the output swing of the phase splitter by its input swing:

$$A2 = \frac{V_{K2MAX} - V_{K2MIN}}{V_{A1MAX} - V_{A1MIN}} = 0.91 \qquad (2)$$

You can approximate the overall distortion at maximum excursion by using the following second-order harmonics formula:

$$\%H_{\text{2nd harmonic}} = \left[\frac{V_{K2quiescent} - \left(\frac{V_{K2max} + V_{K2min}}{2}\right)}{V_{K2max} - V_{K2min}}\right] \times 100\% \approx .1\% \qquad (3)$$

The excellent linearity of the ECC99 and the optimization of the circuit are the first contributors to this very low level of distortion. Being a special case of the cathode degenerative triode, the Concertina also benefits from current feedback, which also greatly improves the linearity.

FURTHER THOUGHTS ON DESIGN

So far, I have calculated the gain and distortion of the phase splitter based on the static loadline, which only reflects the DC conditions of the circuit. Normally, this exercise should be done using the dynamic loadline to obtain the utmost accuracy.

Under AC conditions, when a signal is applied to the grid of the phase splitter, the grid resistors R_g of the power tubes act as a load on the phase splitter and become coupled in parallel with R_{A2} and R_{K2}. While the quiescent point remains the same, the loadline rotates around point Q giving a somewhat steeper slope[13, 14]. However, when the grid resistors are more than ten times the value of the load resistors, the difference between the DC and AC conditions is small enough to be neglected. In my design, I set R_g to 220kΩ (15 times) which decreases the total dynamic load to 28.1kΩ as compared to 30kΩ. The incidence on the slope of the loadline is so negligible that it leads to a rotation of less than 0.5°, making AC evaluation futile.

The Concertina is sometimes disliked by DIYers because of the inherent discrepancy that is found between the output impedance of the anode and the cathode. Though the output impedance is different for the two output legs, the balance will remain unaffected at either low or high frequency as long as the total effective impedance of the anode leg is equal to that of the cathode leg[15]. To meet this condition fully, the stage following the phase splitter must operate in class A at all times, as you find

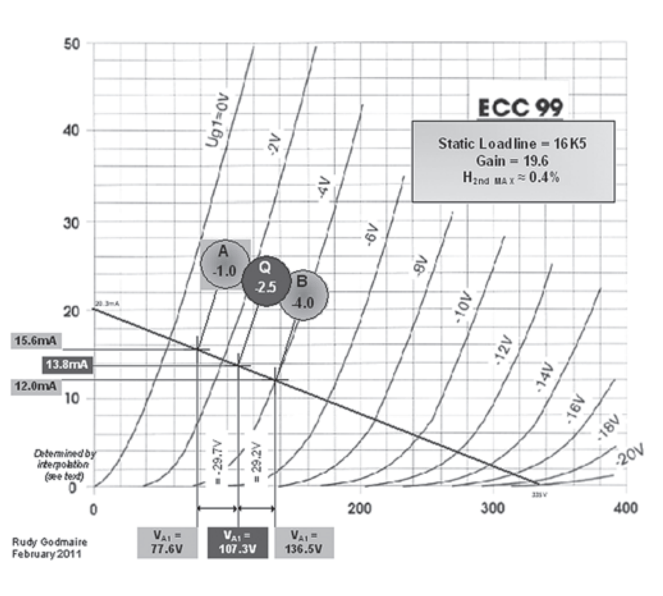

FIGURE 6: Voltage amplifier loadline.

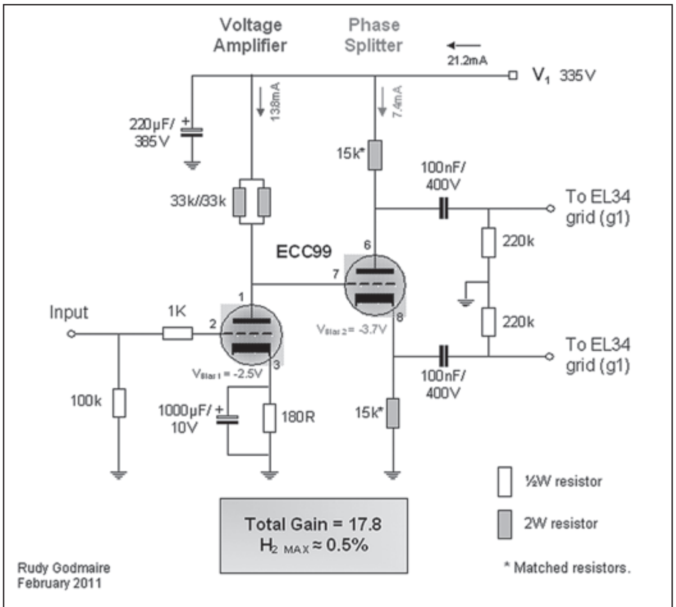

FIGURE 7: The final circuit.

in single-ended topologies. But because my amplifiers use a push-pull output stage, the EL34s will operate in class AB, which means that they will switch from class A to class B over a certain threshold.

Under class B loading, the output impedance at the anode will approximate R_{A2} while the output impedance at cathode will be below 1kΩ. It then becomes evident that lower values of R_{A2} will help mitigate this imbalance. Because the low r_p of the ECC99 makes it possible to select a lower load resistance than in the classic designs using the ECC82, the phase splitter will benefit from an improved performance under all circumstances. From that perspective, the 6H30pi makes an exceptional contender for the Concertina.

Another way to further mitigate the eventuality of an imbalance is to design the output stage in a way that maximizes its class A operation. Using a moderately high tension for the output tubes as in my VDV amplifiers (V_0 = 365V) promotes class A operation because the plate current in any specific valve flows most of the input cycle. Higher values of high tension would increase the output power of the amplifier but at the expense of reduced class A operation.

The absence of amplification of the Concertina is another aspect that is often considered a weakness. While this might be true in some circumstances, it is not a problem in my VDV amplifiers because the voltage amplifier stage preceding the split-load phase splitter provides plenty of gain to drive the EL34s to clipping.

VOLTAGE AMPLIFIER

Going backward again brings me to the input stage; namely, the voltage amplifier. In **Table 1**, I already established that the output swing of the voltage amplifier (ΔV_{A1}) would be ±29.7V. As mentioned previously, you should try to design the circuit so that the signal current in the voltage amplifier is equal and opposite to that in the phase splitter. Provided that both stages use the same type of triodes—as is the case here—this condition will be met when:

$$(4)\quad R_{A1} = \frac{R_{A2}}{A_2} = 16\,484\Omega$$

You can obtain this value by paralleling two 33kΩ resistors which gives a value of 16k5Ω. **Figure 6** shows the loadline of the voltage amplifier based on such load resistance. Note that be-

TABLE 1: DETERMINING POINT B BY INTERPOLATION

$V_{bias\ 2}$ (V)	i_2 (mA)	V_{K2} (V)	V_{A1} (V)	ΔV_{K2} (V)	ΔV_{A1} (V)	Operating Point	Interpolation Factor
-1.0	9.2	138.0	137.0	27.0	29.7	A	
-3.7	7.4	111.0	107.3	0	0	Q	
-6.0	5.9	88.5	82.5	-22.5	-24.8	B'	
-6.46	5.6	84.1	77.6	-26.9	-29.7	B	22.8%
-8.0	4.6	69.0	61.0	-42.0	-46.3	B"	

Note: $V_{K2} = i_2 \times R_{K2}$ and $V_{A1} = V_{BIAS\ 2} = V_{K2}$

cause the input impedance of a phase splitter is very high, I only rely on the DC conditions to make all my evaluations for the voltage amplifier.

Next, you must determine the quiescent point (point Q) of the voltage amplifier and the corresponding value of R_{K1}. Looking at **Table 1** again, you know that V_{A1} must be at 107.3V. Projecting this voltage on the loadline shows that point Q will be at -2.5V, which leads to a quiescent current of 13.8mA. From there it becomes easy to calculate the cathode resistor R_{K1} the usual way by simply dividing the bias voltage by the quiescent current. The resulting value of 181Ω can be closely matched with a 180Ω resistor.

Applying a voltage swing of -29.7V from the quiescent point determines point A on the graph (77.6V). Luckily enough, point A happens to be at a grid voltage of -1V, exactly where I wished it would be. In reality, this did not happen by chance. Three iterations were necessary to reach this optimized value (see sidebar).

Because the input signal coming from the source is undistorted, I know that point B will be situated at -4.0V. Projecting point B on the voltage axis shows that it will be at 136.5V, a variation of 29.2V relative to point Q. Comparing this result obtained from the graph to **Table 1** reveals that the voltage amplifier has some distortion, which you can calculate using formula 3:

$$\%H_{2nd\ harmonic} = \left[\frac{V_{A1quiescent} - \left(\frac{V_{A1max} + V_{A1min}}{2}\right)}{V_{A1max} - V_{A1min}}\right] \times 100\% \approx .4\%$$

You only need to determine the gain of the voltage amplifier. Referring to **Fig. 6**, you find the gain by dividing the output swing of the voltage amplifier by its input swing. Therefore:

$$A1 = \frac{V_{A1max} - V_{A1min}}{V_{g1max} - V_{g1min}} = 19.6 \qquad (5)$$

TABLE 2: MEASURED THD

VIN		THD	
(RMS)	(±)	(dB)	(%)
0.35	±0.5V	-60.0	0.1
0.71	±1.0V	-53.5	0.2
1.06	±1.5V	-50.5	0.3
1.42	±2.0V	-48.0*	0.4*
1.77	±2.5V	-44.5*	0.6*

*FFT analyzer reveals very significant high-order harmonic content due to positive grid current

GLOBAL CIRCUIT PERFORMANCE

Overall, both stages will theoretically exhibit a global second-order harmonic distortion of approximately 0.5% for input voltages of ±1.5V, which will be sufficient to drive the EL34s to clipping. If you compare **Figs. 4 and 6**, you will also notice that the current variations in the voltage amplifier are equal and opposite to those in the phase splitter.

Out of curiosity, I measured the total harmonic distortion (THD) at the anode of the phase splitter using my HP 339A distortion analyzer for various input voltages. The results are presented in **Table 2**[16]. I was very pleased to find that the circuit performs better than I anticipated. When the input signal is set to ±1.5V, the THD reaches a mere 0.3%. This is especially good considering the fact this result is based on THD instead of solely second-order harmonic distortion. Examination of the harmonic content through a FFT analyzer revealed that the harmonics are mainly of the second order for all input levels up to ±1.5V. However, pushing the circuit into the positive grid current area completely changed the pattern of harmonics, causing a very significant increase of third and higher-order harmonics.

I stated earlier that my designs benefit from plenty of headroom. In my VDV amplifiers, the EL34s clip when the excitation of their control grids exceeds ±27V. It is therefore unlikely that the input circuitry will be asked to provide more amplification. Should this happen, the input tube could easily swing well above points A and B. I know, however, that this will lead to higher levels of distortion mainly because of the positive grid current that will progressively occur above -1V. But whether the amplifier distorts because of this phenomenon or because the EL34s clip does not make much difference; it distorts!

Figure 7 shows the final design of the input circuitry developed in this article. This optimized circuit can be utilized as is in the VDV-6040 and the VDV-3070 amplifiers proposed by Menno van der Veen in his book. Using the aforementioned methodology, you may design an optimized circuit for any amplifier having a different voltage swing requirement for the output tubes. For those

using EL84 output tubes, I suggest you try the 6H30pi. In this particular application, I believe this super triode will be an exceptional choice.

Whatever type of input tube you use, I strongly recommend you purchase a matched pair with matched sections for this particular design. This is the best way to ensure excellent balance between both channels, even as tubes age.

I have presented my VDV amplifiers as if they were using a single push-pull output stage. In fact, these are VDV-3070 amplifiers that use a double push-pull topology. Should you use a parallel configuration as I do, you simply double the value of R_g and halve that of C_c in **Fig. 7**. Aside from this, the rest of the circuitry is identical. In both cases, the suggested values will give a -3dB cutoff frequency of 7Hz.

CONCLUSION

Simple tube circuits often benefit from excellent musicality. When these are optimized with care and utilized well within their limits, high fidelity rewards the diligent designer. In this respect, the methodology presented in this article should help you get the very best from your designs.

Should you wish to design your own circuits with the ECC99 using my methodology, I invite you to download my spreadsheet from the *audioXpress* website as mentioned in the sidebar. ***aX***

DRAWING LOADLINES

The old-fashioned way of working with loadlines means that you print a few copies of the plate characteristics, take a ruler and a pencil, and draw your loadlines.

I prefer to use drawing software such as Microsoft Visio. I simply cut the chart from the PDF document and paste it in my drawing software. From there, I can easily draw any loadline I want over the image, and save them separately. One significant advantage with this method is that you can zoom in to make more accurate evaluations. Alternatively, you may also use the drawing features provided by MS Excel. ■

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9. **Mullard EL34 datasheet**, available at http://www.mif.pg.gda.pl/homepages/frank/.

10. The grid bias voltage V_{g1} may also be extrapolated from the pentode characteristics when the triode mode plate characteristics are not available. See my article "V_{g1} Correlations," *audioXpress*, March 2009, p. 32.

11. For further details, see reference #6, chapter 7, section 5 (v).

12. It is possible to calculate the distortion up to the fourth harmonics by using Espley's method. However, because third and higher harmonics are usually negligible with audio tubes, it is usually acceptable to evaluate only the second-harmonic content to figure out the distortion of the circuit. See reference #7, section 4-9 for the description of Espley's method.

13. It is recommended to use a grid resistor R_g that is at least four times the value of the load resistance of the preceding stage. This maximizes the output voltage while minimizing the intermodulation distortion. However, this value must not exceed the maximum stated in the power tube datasheet. See reference #6, chapter 12, section 2 (i).

14. Because of the reactive components of the circuit, the real dynamic path of operation would be an ellipse whose center would be point Q. Realizing such a graphical analysis would be very complicated and this is why it is customary to assume that the load is purely resistive, which enables you to use a straight line to approximate the behavior of the circuit under AC conditions. For further details see reference #6, chapter 2, section 4 (vi); reference #5; and also reference #7, section 4-6.

15. For a detailed explanation, see reference #6, chapter 7, section 2 (ii) B; reference #3, chapter 6, pp. 405-407.

16. The measurements were made without the EL34 output tubes to avoid any distortion coming from their clipping and from capacitor blocking.

1, 2, 3. . . FINISHED!

Implementing the methodology I propose is truly easy. The following hints will further help you optimize your designs in less time than it takes to listen to a CD!

1. Determine the high tension you will use for your input tube. For better results, V1 should range between 325V and 400V; the higher the tension the better.

2. Design the entire circuit as described in the article. Start with a load resistance RK2 that is approximately seven times the value of rp, so 16k.

3. Answer the following question when your design is completed: What is the grid voltage Vg1 of the voltage amplifier at point A?

If it is -1.0V, you are finished!

If it is lower than -1.0V, say -1.4V, redo the design of the circuit using a lower load.

If it is higher than -1.0V, say -0.6V, redo the design of the circuit using a higher load.

Hint: For each deviation of 0.1V of grid voltage, adjust the load by a factor of 0.2rp. If a third iteration is required, extrapolate the value from the first two results.

For your convenience, I have prepared a Microsoft Excel spreadsheet that will assist you when you design your own circuits with the ECC99. This tool, which requires only minimal skills with MS Excel, proposes a Design Protocol that summarizes my methodology in 24 simple steps. The file is available for free download at www.audioxpress.com/magsdirx/ax/addenda/media/ecc99designer.xls. ■

Harman-Kardon HD 990 CD Player and S/PDIF Input DAC

By David Rich

Dimensions: 2.5″ × 17.5″ × 13″
Weight; 8.6 lbs
List Price: $699
Street price: $600
Coax and optical S/PIDF inputs support data rates up to 96k Sa/s (samples/sec) and digital input word widths from 16 to 24 bits.
Balanced and single-ended analog outputs. No DSD (Direct Stream Digital) playback support. Plays CD layer of an SACD.

This unit's performance and features are unique for its price point. While it has the look of a CD player in the $600 price range (**Photo 1**), the back of the unit—with its XLR-balanced outputs—reveals something very different. In addition, there are extra sets of coax and optical S/PDIF plugs marked as input. Input from what? Perhaps a computer or other electronic component that can stream high-resolution music files or stream data directly from an Internet radio station. You can activate the digital inputs with a source switch on the remote control.

Why use these instead of the sound card on your computer? Because the electronics of the HD 990 are state-of-the-art, as I will discuss shortly.

The S/PDIF inputs are connected to a Cirrus Logic 192k Sa/s Digital Audio Interface Receiver, but the inputs only accept 96k Sa/s inputs on the HD 990. The unit mutes if a digital signal is sampled at a rate of 192k Sa/s, as is the case for other units I have tested with S/PDIF receivers limited to locking to only a 96k Sa/s maximum signal. The audibility of a sampled signal at 192k Sa/s over one at 96k Sa/s is debatable, and most computers or other sources capable of playing a 192k Sa/s sampled file should have an option to limit the sampling rate to 96k Sa/s so it can be played on the HD 990. When moving from a CD rate sampling rate of 44.1k Sa/s to 96k Sa/s, an audible difference is much more likely and is clearly measurable.

Many computers do not have S/PDIF outputs and instead send digital music down the USB output. Relatively inexpensive interface units between USB and S/PDIF are available to handle sampled signals up to the 96k Sa/s rate. I did not use these units with the HD 990. With its S/PDIF inputs, acquisition occurred without any clicks or pops.

The digital outputs have a maximum sampling rate of 44.1k Sa/s, meaning the higher rate-sampled signals are converted to 44.1k Sa/s. Harman reports the outputs are designed to interface with recording devices which are typically limited to 44.1k Sa/s. It makes little sense to take audio off the unit digitally given all the sophisticated electronics in this unit used to convert the digital signal to analog.

The unit is equipped with the Harman proprietary HRS Link output that works with some Harman AVRs and the matching HK 990 stereo integrated amplifier with DSP. It is bidirectional (four wires) with the system clock for the DACs in the amplifier sent to the HD 990. This forces the rate of data transmission from the HD 990 across the HRS link to the integrated amplifier or AVR to match exactly, eliminating the need for an asynchronous data converter on the amplifier side. The HD 990 is the lowest-priced CD player in the Harman line with the HRS link; a digital connection with HRS link may make sense to take advantage of functionality, such as room correction, that operates in the digital domain of the

PHOTO 1: The HD 990, Harman-Kardon's stylish CD player/DAC.

Harman AVRs and HK990 integrated amplifier.

Inside the Unit

Without a service manual, I can only provide direct observations of the two PC boards on the chassis. One is the power supply, the other is the main board with the electronics. The power supply is a switching unit increasingly common in mid- and low-priced CD and DVD players. The board is 3.5 × 7″. Analog power supplies are ±14V at the output of the switcher. Linear regulation for the DACs and analog operational amplifiers is on the main board, which contains the electronics for the CD drive (no electronics under the drive), the DSP chip and associated memory, the DACs, analog electronics, and digital audio interface chips.

Digital data from the CD or the digital inputs of the HD 990 are passed to the DSP chip, which is a 32-bit Analog Devices Blackfin DSP that suppresses jitter and does digital reconstruction filtering. This DSP chip is absent from most CD players and is rather expensive by itself, not to mention the higher build price when the cost of the memory chips for the DSP is included.

The DSP suppresses jitter using an Asynchronous Sample Rate Converter (ASRC). As the name implies, signals with different input and output clocks are reconciled with this signal processing block. If the incoming signal has an input clock with jitter, the ASRC will re-clock the signal to a low jitter clock (the output clock) from the crystal oscillator on the main board. The presence of an ASRC is especially important if the unit operates with an external S/PDIF signal.

The next operation the Blackfin DSP performs is digitally filtering the signal after it performs the sample rate conversion. The process of passing through the digital filter produces a digital datastream from the DSP chip at 384k Sa/s at a 24-bit depth. The DSP has not produced the same amount of information available from an analog music signal actually recorded at 384k Sa/s rate and a 24-bit density. Instead, the DSP has suppressed part of the spectrum of the sampled signal.

For a band-limited 20kHz analog signal sampled at a rate of 44.1k samples/sec, the spectra appear around 0Hz (desired signal) as well as from 24.1kHz to 64.1kHz and then repeat around multiples of 44.1k. The repeating spectra are the stop-band images that result from the sampling process. The suppression of the stop-band images allows the sampled signal to better represent an analog signal sampled at 44.1kHz. Information lost as a result of band limiting the signal before the sampling process cannot be restored. The signal-to-noise (SNR) ratio of the 16-bit quantization also cannot be lowered, though the standard practice of adding dither before the 16-bit ADC ensures the quantization residuals are noise, not correlated harmonics.

Digital filters can be designed in the time domain, the frequency domain, or a combination of both. Harman-Kardon indicates the filter uses a reconstruction process called Real-Time Linear Smoothing (RLS III), but provides no details.

Most CD players lack an ASRC and leave the digital filtering to the DAC chip, which has internal digital filters. The rejection of the stop-band images from the sampling process would not be as high if the filtering is done by the DAC chip because there is only limited silicon space for DSP filtering on the DAC chip compared to the space on a dedicated DSP chip.

The DACs on the HD 990 are top-flight Analog Devices AD1955A chips. Although the AD1955 has two DACs on each chip for stereo, the HD 990 uses one per channel in a balanced configuration for reduced noise and distortion. The balanced digital inputs are developed on the Blackfin DSP. In mono mode the AD1955A has an equivalent worst-case noise floor of 19-bits (A-weighted). Current to voltage conversion and analog reconstruction filtering, respectively, are performed by an Analog Devices OP275A and a Texas Instruments OPA2134A dual operational amplifier. The output stage is fully discrete to drive low impedance balanced loads. Two discrete circuits for each channel are on the board to drive the balanced outputs. I saw no signs of a DC servo to reduce the number of inter-stage DC blocking capacitors.

I cannot measure a unit with the HD 990's performance. I refer you to the Australian site avhub.com.au for a complete set of measurements as a CD player. Steven Holding's results are exemplary, as you would expect from Analog Devices' highest grade DAC surrounded by quality op amps. The noise floor is remarkably low with no signs of any power supply noise.

You can see the presence of the Blackfin DSP in the measurement of the worst-case stop-band image attenuation for a 44.1k sample/sec signal (slightly above 24kHz for a 20kHz sign wave sample at the 44.1k Sa/s rate). The stop-band image is suppressed by over 100dB. I normally would not report jitter results because I believe its significance on audio quality is overstated. However, Mr. Holding reports they are as low as he has measured. This verifies the algorithms for jitter attenuation have been well chosen for the Blackfin DSP chip.

Ergonomics and Performance

The styling of the unit is reminiscent of the Harman Nocturne receiver introduced 45 years ago. When the power is off, nothing is visible. The downside of the ultra high-tech look is the tedious front panel controls. The switches to operate the unit are pencil-thin at the center left of the unit with the exception of the open/close button at the right side of the CD loading tray. The buttons are the same color (black) and

identically sized, offering no distinction by function.

To make matters worse, lettering is small and with little contrast. The remote has larger buttons of varying sizes for the main control features, but the small letters and poor contrast remain. Secondary controls are all the same-sized round buttons. The critical source/select button is especially hard to identify.

The display shows the sampling rate for S/PDIF inputs, CD text information, or ID3 tag information for MP3 files. The graphics are large with good contrast. A DVD player used as a CD player replacement would display none of this information. Instead, the text would be shown on the TV connected to the DVD player; in a stereo music setup where no TV would be found, this information would be lost. Unfortunately, the instruction manual (which you can download from the company's website www.harmankardon.com) is only 16 pages thin and does not highlight how the display varies across the unit's operating modes.

The unit had a problem with the wider gaps in the Pierre Verany error correction test disc, but it played CDs without any click or skips that tripped up players that performed better with the Verany disc. This left me questioning the value of the test disc, not the performance of the player.

I heard no degradation in sound quality from the unit. Given the parts quality and measured performance, this comes as no surprise. Others might hear some difference between the HD990 and higher priced units, but I would be very surprised if they found fault with this unit.

A Comparison

The Cambridge Audio Azur 840C offers a good comparison. It shares the same Blackfin DSP and a pair of AD1955A DACs. It also has S/PDIF inputs and balanced outputs. At $1800, it is three times the price and has significant build quality differences.

The power supply is traditional with the transformer at 60Hz. The unit is twice as heavy owing to more substantial sheet metal and the larger power transformer. The disc transport is perhaps more robust and the electronics associated operating the transport are different. The ergonomics at the front panel and the remote are substantially better.

Cambridge supplies a 14-page white paper on the novel jitter-reduction and digital-filtering algorithms encoded in the Blackfin DSP. As with a computer, hardware performance is often secondary to the software loaded on it. Indeed, the distinguishing feature of four- and five-figure CD players and DACs is often the algorithms in the DSP. Harman offers no information on the signal processing, though the measurements cited above leave little doubt the performance is improved compared to a CD player that does not have an external DSP chip on the board.

The Cambridge Audio Azur 840C S/PDIF input will accept a 192k Sa/s signal that the Harman HD 990 cannot. The 840C also transfers data at the external S/PDIF input with a 24 bit depth to the DSP. Harman personnel could not confirm whether the HD 990 sent 24 bit data to the DSP or truncated it to 16bits.

Conclusion

Short and simple: the quality of the electronics in this CD player and the resulting analog performance are nearly impossible to beat at this price. The state-of-the-art electronics with external S/PDIF data makes an already highly desirable unit even more so by combining what are normally two different components into one box. I was bothered by the uniformity of the front panel buttons and the minimal contrast of the lettering on the front panel and remote, though through repetitive use you soon develop spatial awareness of the functions. *rr*

Notes on Notation

The sampling rate defines the number of samples per second taken from a continuous time wave to produce a discrete signal. I used the notation of the sampling rate as samples per second or in the abbreviated form Sa/s.

It is typical to specify the sampling rate in Hz, but then you may become confused by what is the frequency of a continuous time signal (cycles per second) and a sampled signal (samples per second).

A CD player data is quantized to 1 of 65536 levels which can be represented by a 16-bit word. The data rate coming from a quantized and sampled signal is the bit rate and is measured in bits per second.

For a CD per channel the bit rate is (16-bits/sample) (44.1k samples /sec) or 705.6k bits per second (often abbreviated BPS). For stereo the CD data rate is two times 705.6k BPS, which is 1411.2k BPS. This stream of data per second is often called the data rate.

A signal sampled at 96k Sa/s with a quantization of 16777216 levels (24 bits) has a bit rate in stereo of 2 * 96k Sa /s * 24-bits, which is 4608k BPS.

A digital data rate coming off a CD may be too fast to stream over the Internet or produce a file too big to store. The data rate can be compressed using a processor such as the MP3 encoder. For example, an MP3 file may stream at 128k BPS. Comparing the data rate of a CD to the MP3 rate, you see the signal is compressed by 1411.2k BPS/128 KBS, or a factor of 11.1, which is the compression ratio having no dimension. —DAR ■

By Jeff Macaulay

Efficient **100W Class A** Amplifier

Experience good class A sound with this author's circuit design.

My first introduction to class A amplifiers occurred in 1974. I had read an article by G.A. French in the now-defunct (sadly) *Radio Constructor* magazine that described how you could produce a class A output stage by combining a constant-current source with an emitter follower. This was just within my design skills at the time and so I produced my own version (**Fig. 1**).

Although primitive, the sound was a revelation. So smooth, with huge bass extension and no sense of listener fatigue. Unfortunately, there was a downside. You could have fried eggs on the heatsinks—or the mains transformer come to that—and power output was rather too limited to cope with the fairly low efficiency speakers I had at the time.

However, the sound set a benchmark for me for what a good amp should sound like and set me off on a quest for an efficient class A amp. Now I know that this sounds like an oxymoron, but for most of the time class A amplifiers, especially single-ended ones, simply dissipate unnecessary power. Listening to music is not, as a rule, done with an amplifier delivering full output. In the nature of things most musical information is at a low level punctuated with sudden and unpredictable peaks. The result is that class A amps just generate unnecessary heat, because the amps' quiescent current is fixed at the peak current required.

In fact, the inefficiency assigned to class A amps in general and single-ended designs in particular is better characterized as a design fault rather than an inevitable feature. It took me only a few months to work out a way to dynamically alter the constant-current source of my amp so that it stayed in class A but responded to signal levels and had a low quiescent. The fly in the ointment was that the circuit was rather touchy and difficult to adjust, so I abandoned the idea.

SHELTER FROM THE STORM

There matters stood until one rainy day in 2000, while trying to avoid the weather in a shop doorway, I suddenly had a brainstorm. On one of those rare occasions I saw the circuit complete in my mind. Luckily, I had a pen handy and hastily drew the circuit on the back of a till receipt.

Frankly, I looked at my circuit in disbelief. It was just too simple to work, although I couldn't fault my own logic. As soon as I got home I simulated it on Electronic Workbench™. To my considerable surprise it worked as I had supposed. In short, it operated in single-ended class A without requiring a high quiescent current.

More important, the distortion produced was mainly second harmonic, just like a conventional class A stage. It also sounded the same as the traditional circuit. After some more work on the circuit, I took out a patent on it, GB2374220.

Every year since I've sworn that I would get out and exploit it, but every year I've paid my patent continuation fees and put it on the back burner. This year, after another hefty payment, I decided enough was enough and so I've finally got around to writing it up.

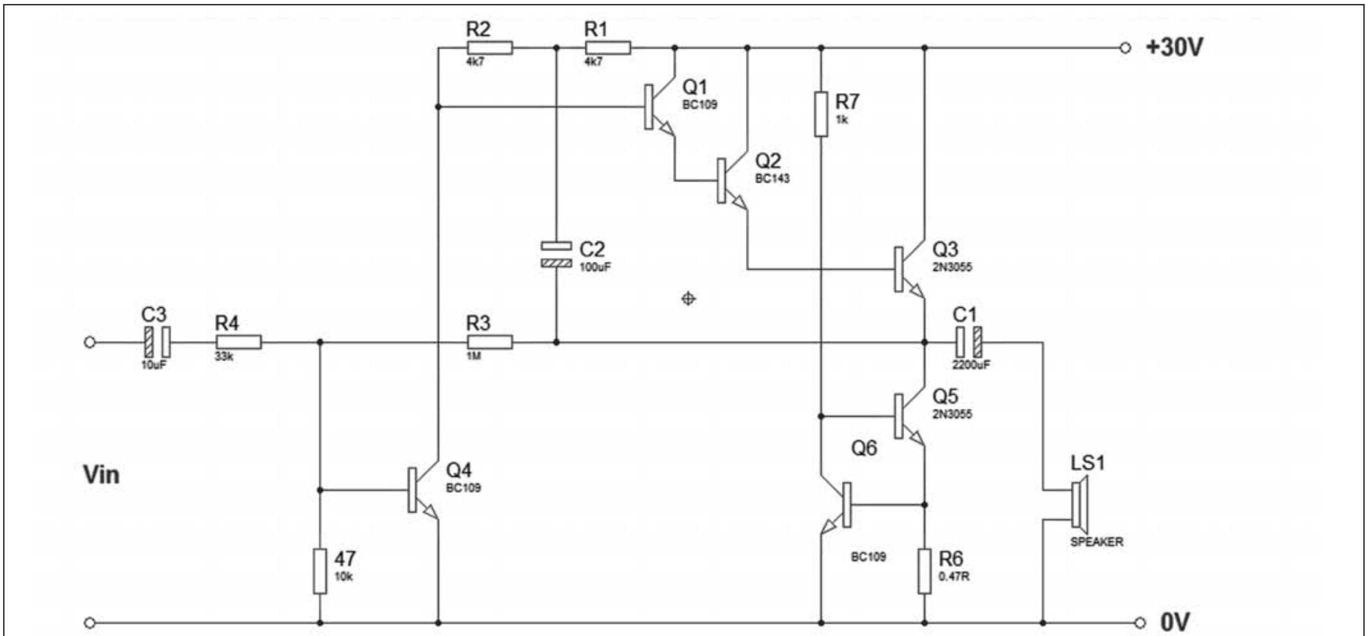

FIGURE 1: The author's original class A amp.

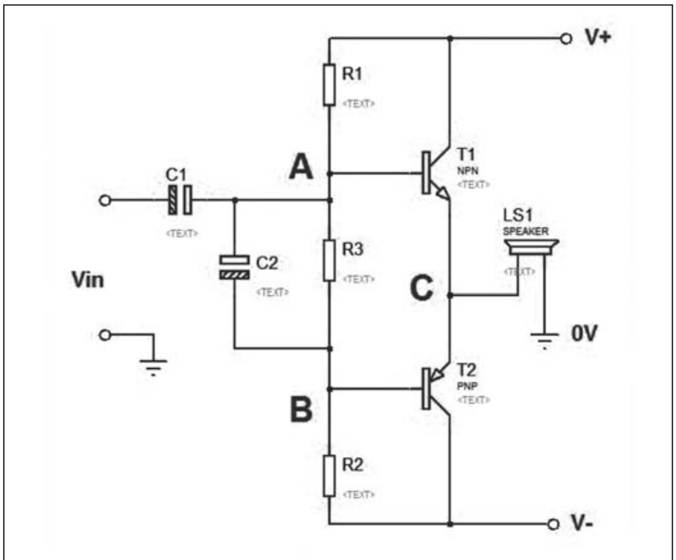

FIGURE 2: Class B push-pull output.

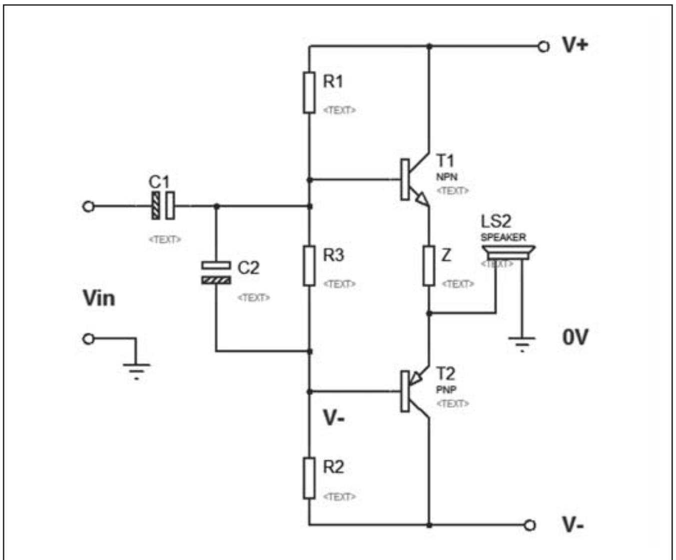

FIGURE 2A: Impedance between the output devices.

CIRCUIT OPERATION

To understand the genesis of the circuit, you must return to the basics. The simplest class A output stage is the single-ended type. Here, a single device conducts through the entire signal cycle. To do this, the quiescent current must be equal to the peak load current required. As a result, the amplifier has a maximum efficiency of 25% and, in practice, seldom manages more than 12%.

Compare this to the class B push-pull output in **Fig. 2**. Here, in its simplest form, it comprises two complementary transistors. R1, R2, and R3 bias the transistors just into conduction in the absence of an input signal, causing a small quiescent current to flow through T1 and T2. The output at the emitters of T1 and T2 is directly connected to the load.

The bias voltage A,B is the sum of the bias voltages for T1 and T2, CA and CB, respectively. On positive signal excursions T1 turns on, increasing the CA bias voltage while reducing the CB bias for T2, which then turns off. On negative going excursions, the position is reversed with T2 turning on and thus turning off T1. The result, push-pull operation, produces large amounts of odd-order distortion, which sounds harsh (crossover distortion) as the signal forces the transistors to turn on and off. Even-order harmonics are largely canceled out by the circuit.

Some thought will show that if either T1 or T2 could be prevented from turning off during the signal cycle, you would have instead a single-ended class A stage combined with a compliant current source. In other words, an efficient class A output stage without the giant heatsinks.

FINDING Z

While sheltering in the shop doorway, I realized that all I needed to do to produce such a circuit was to include a single impedance, "Z," between the output devices as shown in **Fig. 2A**. What happens then? Well, if the impedance is properly chosen, the output stage is unbalanced to such an extent that T1 (in this case) won't turn off until T2 reaches its maximum negative excursion; just like a conventional single-ended output stage, if T2 were replaced with either a resistor or constant-current stage. No crossover distortion is generated, but even harmonic distortion becomes prevalent in the output signal, again exactly what you would expect of a single-ended output stage.

Paradoxically, single-ended class A operation generates more measurable distortion than class AB stages—the open loop THD is on the order of a few percent—with this circuit. However, you can easily reduce this to minute levels with a dash of negative feedback. Much more important is that the basic class A sound is there both when used open or closed loop.

It is not such a problem to calculate the value of "Z" as you might first expect. It is a characteristic of transistor junctions that the base emitter voltage increases by about 60mV for every tenfold increase in current flowing through it. This relationship holds true over six orders of magnitude.

You can use this to calculate the value of the impedance required. In principle, all you need to know is the change in Vbe over the current range of the output stage. The impedance is then equal to that voltage divided by the quiescent current.

For example, take the current design. The spec requires a quiescent current of 50mA, I_q. You can calculate the peak current, Imax, from the maximum power, P, and load impedance, R_L

$$\text{Imax} = 1.41\ (PR_L)^{0.5}\ R_L^{-1} = 4.99\text{A}$$

Similarly, you can find the number of decades of current, x, from the relationship $10^x = \text{Imax}/I_q$, hence $x = \log(\text{Imax}/I_q) = \log(4.99/0.05) = 1.99$ decades.

Then the change in Vbe, V, = 0.06x = 0.119V.

And finally you can calculate Z:

$$Z = V/I_q = 0.119/0.05 = 2.39\Omega$$

Or, $Z = 0.06 \log (1.41\ (PR_L)^{0.5} RL^{-1} I_q^{-1}) I_q^{-1}$

DESIGN CONSIDERATIONS

Although I have used this output stage over the years, I don't, as a rule, design high power amps. Although I realize with the passage of time the definition of "high powered" has drifted up from a 10W output in the early 70s to 100W minimum today. But more power output was only one aspect of the design of this amplifier. I wanted to produce a design that will reliably operate over a large range of supply voltages without modification. Here the balanced design approach pays dividends. I have successfully used this design from ±9 to ±50V power supplies, without the need of adjustment.

Also, when it comes to the output

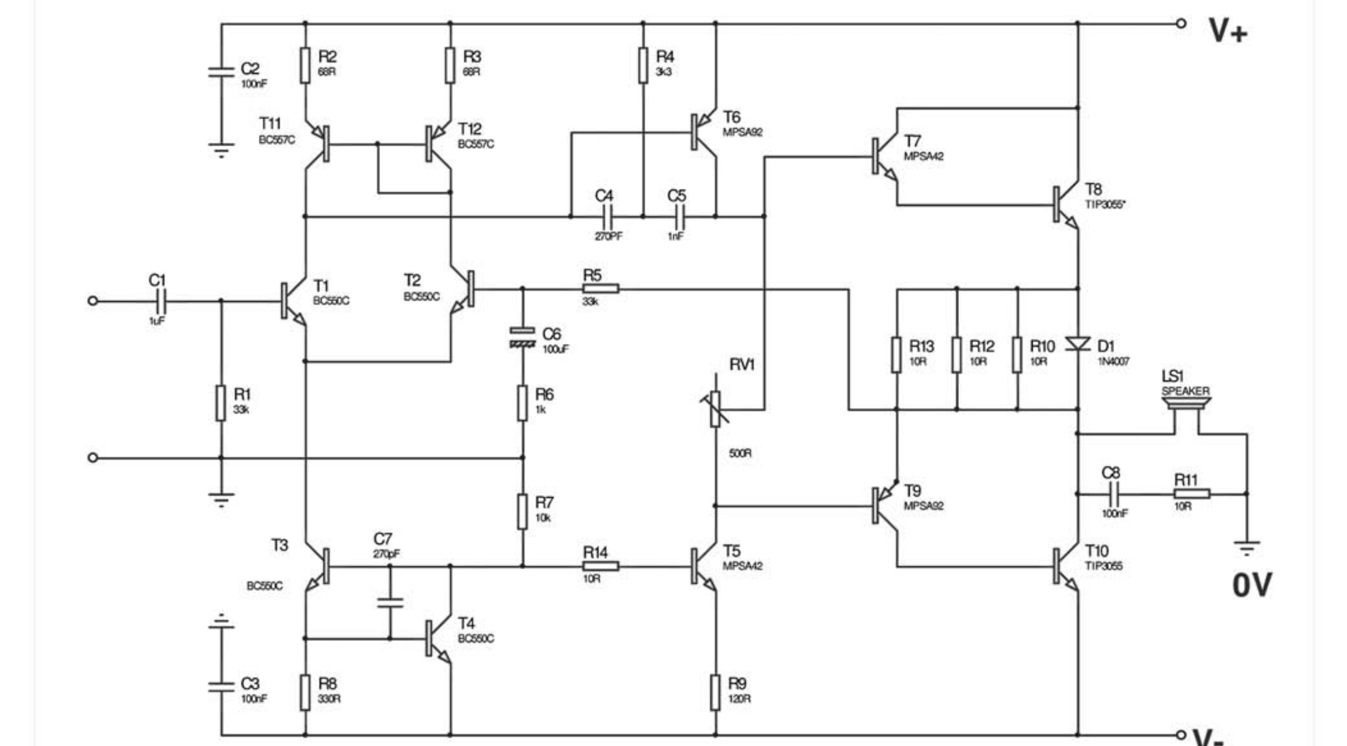

FIGURE 3: One channel of the amplifier.

stage, a compromise must be reached between open loop distortion and quiescent current as determined by the series impedance. This is a matter of judgment and I have come down on the side of high impedance, 3.3Ω, and open loop distortion. Why? Well, one useful result of using a higher impedance is a lack of sensitivity to thermal runaway. In short, you can set the quiescent current by a simple preset; thermal feedback is not required! So, no difficulty setting up. Indeed in my previous designs of this type I have used a fixed resistor for this duty but I decided to allow you something to fiddle with, should you feel inclined.

If you calculate the power rating required of a 3.3Ω resistor in series with an 8Ω load dissipating 100W of audio, you end up with the rather startling result of 82.2W peak! Obviously, this is not a satisfactory result and so you must take measures to shrink this to a reasonable level. This is achieved by the use of a shunt diode wired in parallel with the 3.3Ω resistor. This allows high output power from sensible sized power supplies because the diode conducts the output current once the 0.6V threshold is exceeded. It also actually reduces the distortion of the stage without altering its efficiency whilst the dissipation of the 3.3Ω resistor is reduced to about 50mW.

More important, though, normal global feedback will reduce the distortion generated in the output stage to minuscule dimensions. TINA™ simulation suggests the amp generates about 0.004% THD, mostly 2nd harmonic at 1kHz at any level below clipping.

The devices chosen for the amplifier are also widely available and inexpensive. This is an important consideration; if I had a pound for every time I've used a new highly vaunted device in my designs only to find it becomes either unobtainable or prohibitively expensive after a couple of years, I could probably retire on the proceeds!

Another consideration is the overall stability of a design. Experience has taught me that the safest way to test this is to build a prototype on strip-board, because there are so many more reactive strays on a strip-board than any other method of construction. This design passed this test with flying colors; as I write, I am listening to my Vero-board™ prototype.

Figure 3 shows the schematic of one channel of the amplifier. The other channel is identical. There aren't many surprises here except perhaps the use of double pole compensation which I will discuss shortly. As an overview, the amplifier can be split into four major sections: a long-tailed pair fed by a current mirror, a driver, and, finally, the output stage.

HOW IT WORKS

Input signals are fed into the amplifier via C1. This device performs two separate functions. First, it blocks any DC that might be present with the signal and that might upset the balance of the long-tailed pair, T1 and T2. Such an imbalance would lead to unwanted output offset voltage across the speaker. Second, C1 defines the low-frequency limit of the amp at around 5Hz.

At the heart of most modern power amps is the long-tailed pair differential amplifier; in this design it comprises T1 and T2. T's function is to compare the amp's input signal with the output of the amplifier via the feedback loop. The output of the stage is the difference between input and output which becomes the feedback signal for the amp. Here the usual collector resistors are replaced with a current mirror comprising T11 and T12. A small but significant amount of negative feedback is applied by R2 and R3 which increases the output impedance of the circuit. The use of a current mirror greatly improves the linearity of this part of the circuit.

The driver circuit consists of T6 and its constant-current source T5. T6's base is fed directly from the collector of T1.

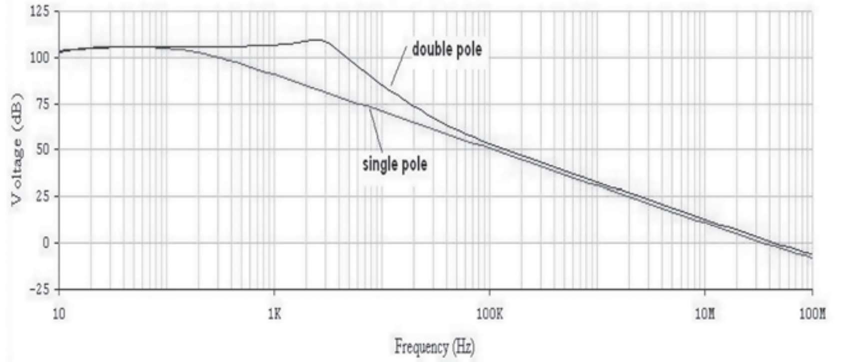

FIGURE 4: Open loop frequency response.

The driver, a high voltage pnp device, is used in the common emitter mode for maximum voltage gain. Because of the high thermal stability of the output stage, bias is obtained by simply adjusting P1. To ensure high open loop gain, T6's collector is loaded by the constant-current source comprising T5 and R9.

Obtaining a large open loop gain for the amplifier is fairly straightforward. More difficult is to use this gain effectively. To explain, in order to maintain stability, the gain of the amplifier needs to fall below unity before 180° phase shift occurs. Due to the inevitable parasitic inductance and capacitance of the circuit, oscillation is guaranteed when feedback is applied unless measures to control open loop gain are applied.

Normally the loop gain would be rolled off at 6dB/octave by connecting a small capacitance across the T6's collector/base circuit. This works fine but better performance can be gained by using a more complex network. Here, double pole compensation is used. C5, R4, and C4 form a second-order filter at low frequencies so that the gain rolls off more quickly initially than a single capacitor would provide.

At yet higher frequencies C5 looks like low impedance to the signal which causes the filter to revert to first-order mode. The advantage is that the turnover frequency of the network can be pushed up while retaining stability. This means that more feedback can be safely applied at the critical mid- to high frequencies, lowering the overall distortion of the amp. **Figure 4** shows the open loop frequency response of the amp with first- and second-order compensation circuitry for comparison purposes. Using the latter, 17dB extra feedback can be applied in the midrange.

T1, 2, and T6 all need to be supplied with constant-current. This is the function of the circuit block comprising T3, 4, and T5 with the associated components R6, 7, 8, 9 and C7. The basic circuit is formed by the directly coupled transistors T3 and T4. T4 controls current through T3 by the base emitter voltage of T4 which appears across R8; R7 completes the circuit providing collector current for T4. To maintain stability, C7 is shunted across T4's collector/base circuit. Because the voltage on T4's collector is constant, this is also used for base biasing T5 providing constant-current for T6.

Having discussed at some length the operating principles behind the output stage, I'll keep the description of this part of the circuit short. T7, 8, 9, and T10 form a basic quasi-complementary output stage. T7 and T8 are configured as a Darlington pair and operate in class A as previously discussed. Negative polarity output current is handled by T9 and T10. The impedance comprises the parallel combination of R10, R12, R15, and D1 which force the output stage to operate in Class A. R11 and C8 compensate for inductive speaker leads.

While on the subject of stability, C2 and C3 are essential. Power supply cabling is prone to rf and noise pickup which can potentially cause oscillation. The final part of the circuit is the feedback loop comprising R5, C6, and R6. The amplifier's gain is set by the ratio of R5 to R6, at ((R5/R6)+1). C6 looks like an open circuit at DC while appearing as a short circuit at audio frequencies. This ensures that the amplifier's gain is unity at DC and thus a low output offset voltage.

As indicated earlier, the amplifier is versatile enough to build on strip-board, although a suitable PCB layout would definitely be better. Whatever method of construction you use, the power devices need to be mounted on a suitable heatsink. A single 1° C/W is adequate for a stereo pair of amps. Don't forget the mounting hardware! Construction should bring few problems. Be sure that the semiconductors and electrolytic are correctly oriented. Caution: the pin-outs of the small signal transistors differ between types.

Last, before you can listen to your new amp you'll need to adjust it. Because the design operates from a large number of supply voltages, a PSU circuit is not provided. Nothing special is required—just a standard ±9V DC to ±50V DC supply. Naturally, the output power depends entirely upon the voltage used. For 100W output the ±50V DC supply will be required. If you only listen to music and not test tones, a 300VA transformer will be more than adequate for stereo use.

To set up the amp, attach the power supply and speaker to the amp, making sure it's disconnected. Adjust PR1 to short out T7 and T9 bases. Check for this on a multimeter set to indicate low resistance.

Now the quiescent current needs setting up. Monitor the voltage across D1 with a multimeter set to clearly show 1V DC. With no input signal applied, switch on the PSU. Nothing should happen! If everything is OK, slowly adjust RV1 for 200mV. Switch off and remove your multimeter. That's it, job done. All that remains is to apply an input signal, sit back, and enjoy true class A sound. *aX*

AMPLIFIER PARTS LIST FOR FIG. 3

RESISTORS	
R1, R5	33k
R2, R3	68R
R4	3k3
R6	1k
R7	10k
R8	330R
R9	120R
R10-R14	10R
RV1	500R
CAPACITORS	
C1	1µF
C2, C3, C8	100nF, 100V WKG CERAMIC
C4, C7	270pF Ceramic
C5	1nF Ceramic
C6	100µF/25V Electrolytic
SEMICONDUCTORS	
T1-T4	BC550C
T5, T7	MPSA42
T6, T9	MPSA92
T8, T10	TIP3055
T11, T12	BC557C
MISCELLANEOUS	
T03P Insulating kits	
Heatsink	
PCB	
M3 Screws	

The Michele **MC Preamp**

Discover how this battery-powered, low-output moving coil tube preamplifier stacks up against conventional AC power units.

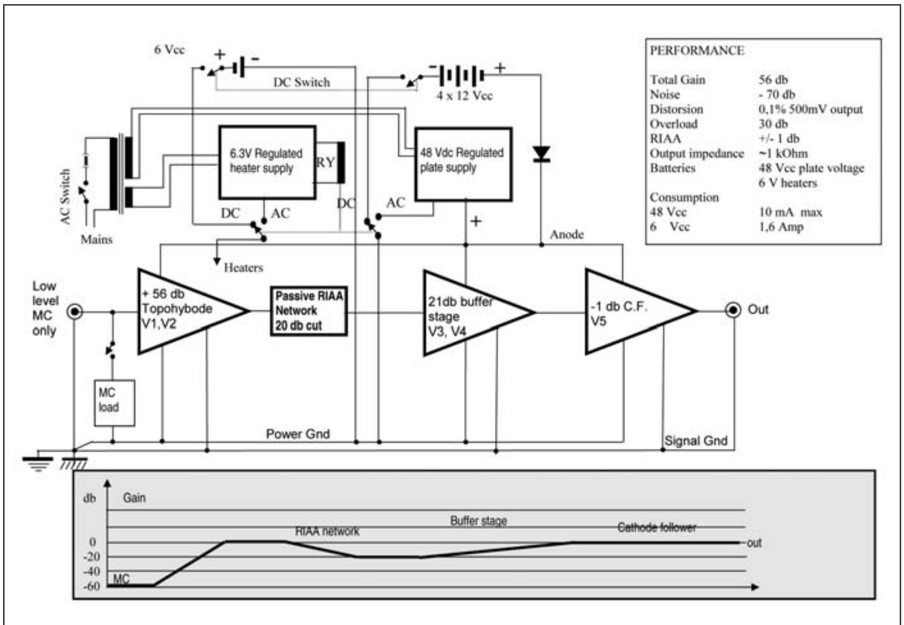

FIGURE 1: Michele MC preamplifier block diagram. The graph at the bottom shows the gain/loss of each stage.

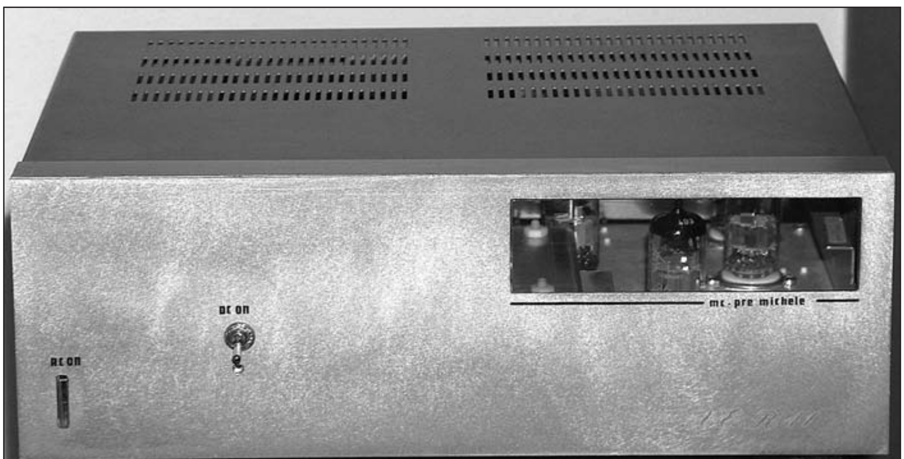

PHOTO 1: The Michele preamplifier.

My purpose in building a battery-powered, tube-based preamplifier was to show the benefits of this type of power compared to a traditional AC power supply with rectification by diodes. This project required the use of a tube capable of working with low plate voltages on the order of 20-30V. I chose a tube I have used in the past for high-frequency amplifiers in cascode circuits, the ECC86/6GM8, whose main features are:

Va (Anode voltage)	25V
Ia (Anode current)	20mA maximum
Vf (Heater voltage/current)	6.3V 0.3A
μ(Amplification factor)	14

My aim, with this project, was to achieve the following:

A. the use of tubes, except in the moving coil input stage
B. zero feedback
C. passive RIAA equalization
D. 48V DC maximum anode, DC voltage (in a totem pole configuration)
E. hi-fi standard performance

CIRCUIT OPERATION

The block diagram in **Fig. 1** shows only one channel. A high gain input stage provides a 56dB gain required for low output MC cartridges (200-500µV). This stage must also have a high signal-to-noise ratio in order to minimize hum and hiss. Its output feeds a passive RIAA equalization network which has a negative gain of 20dB at 1kHz, immediately restored by the next buffer stage. From these, the signal through a cathode follower is applied to the RCA output connectors.

The power supplies section is made up of voltage-regulated circuits both for heater and plate voltages. A relay provides proper voltages to the circuit either from an internal power supply or external batteries. When the pre is fed through an AC outlet, relay RY1 is automatically energized and feeds all circuits by means of internal power supplies. Under this condition, the AC/DC switch does not have any effect.

If mains is turned off and you set the switch to DC, the preamplifier is then fed by the DC voltages supplied by the external rechargeable batteries. (12V/1.3 A/h × 4 = 48 Vcc for plate voltages, plus 6V/12A/h for heaters). The diagram at the bottom of **Fig. 1** shows graphically the gain/loss of various stages.

The construction of this preamplifier requires a thorough knowledge of audio frequency electronics, assembly techniques, and the problems inherent in high sensitivity circuits. ***It is not recommended for inexperienced DIYers.***

DESIGN CHOICES

The whole design resembles, to some extent, a more sophisticated circuit designed back in the 80s by the Italian engineer Bartolomeo Aloia.

Having established a set of requirements (A through E above), it becomes immediately apparent that a conventional cascade of triodes is unable to meet the prerequisite because it is impossible to obtain a distortion of less than 1% unless several dB of feedback are applied to the circuit. This conflicts with requirement B and E, which note that zero feedback and high fidelity requirements are design goals.

Additionally, a high performance circuit must be "fast" (high slew rate) and present a wide bandwidth in order to provide good response to the transients. To achieve these design requirements, I surveyed alternative circuits and ultimately focused on the SRPP (Shunt Regulated Push-Pull) configuration, better known as totem pole. It has been largely used in the past in instrumentation and professional fields as high frequency line driver and voltage amplifier. Its operation has been described several times in aX, and you can also find an exhaustive description at www.tubecad.com/may2000/page2.html.

Finally, as stated earlier, in order to operate with low plate voltage, I had to select a suitable valve. Among limited alternatives, the ECC 86/6GM8 was a good choice because it is still available at a reasonable price.

The totem pole circuit, with the appropriate components values, provides a gain of about 20dB (10 times) with a distortion on the order of <0.1% at 1V output without application of a feedback loop, a high slew rate, and a bandwidth reaching a MHz.

Low-level signals, such as those coming from the MC cartridges, are in the range of 200-500μV at 1kHz and require an amplification factor up to 1000 times (60dB) to reach at least 200-500mV output level needed to drive a line amplifier. A vacuum

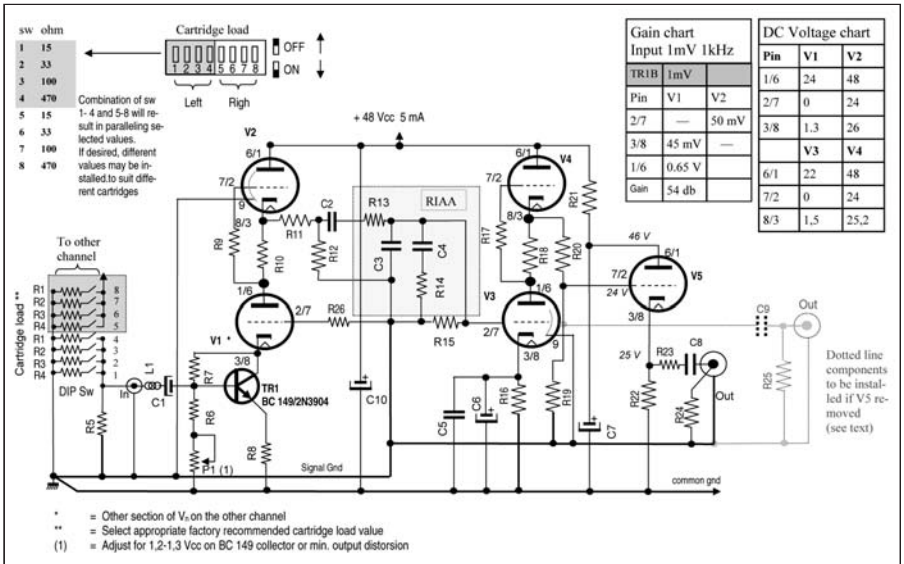

FIGURE 2: Michele MC preamplifier schematic, one channel shown. Tube pin numbering (i.e., 1/6) shows left/right channel pin assignment.

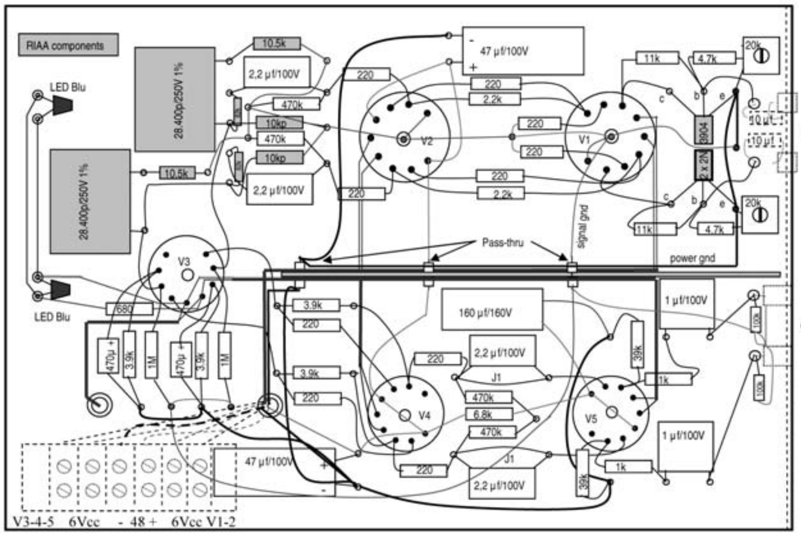

FIGURE 3: Point-to-point wiring arrangement. On my project I always draw by hand a sketch of all connections. This helps reduce errors and improves circuit layout.

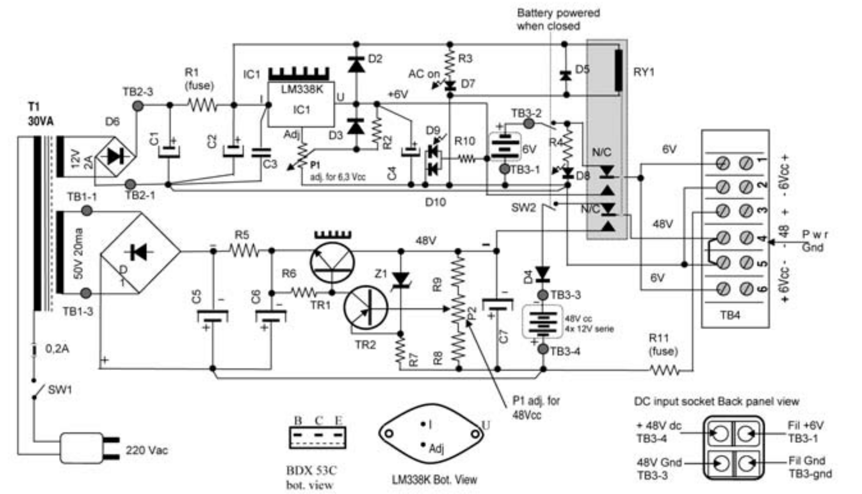

FIGURE 4: Power supply schematic. LM338K must be mounted on a 5 × 10 cm or larger heatsink. 48V and 6V batteries are located outside preamplifier chassis.

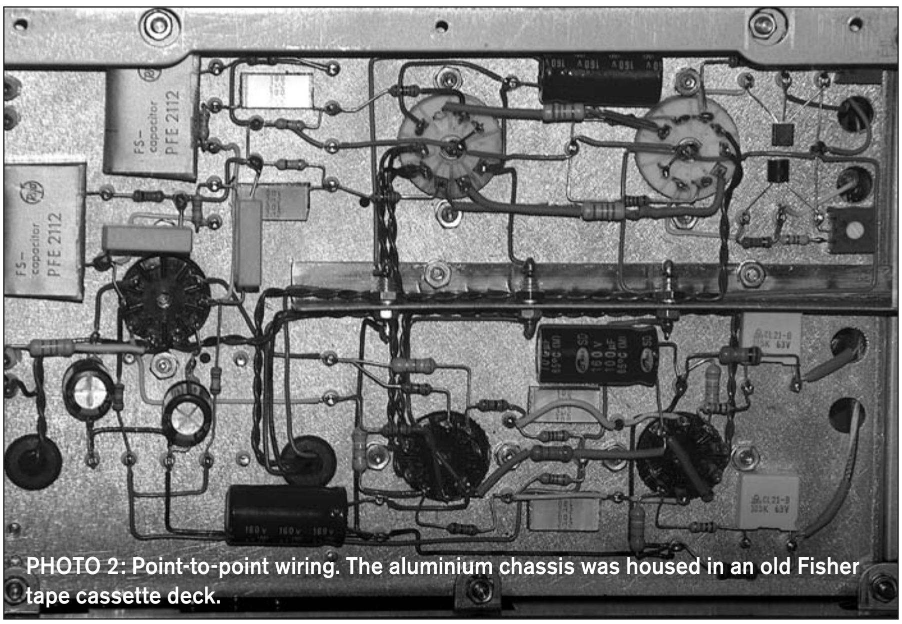

PHOTO 2: Point-to-point wiring. The aluminium chassis was housed in an old Fisher tape cassette deck.

tube circuit can hardly respond to these requirements considering also the intrinsic noise problem introduced by them. A brilliant solution was identified, back in the 80s, by the designer Aloia, who called this particular circuit TOPOHYBODE—an acronym for TOtem POle HYBrid cascODE. It consists of a totem pole in a cascode configuration with an NPN low noise transistor (**Fig. 1 V1**).

The performance of this circuit is outstanding: it exploits all advantages of the vacuum tube and the transistor. With its simplicity and over 30dB overload capability, it surpasses by far any other solution. Finally, a passive RIAA equalization network completes the hard part of this circuit.

CIRCUIT IMPLEMENTATION

Low-level input stages must be rather quiet in order to maximize the signal-to-noise ratio necessary for good performance. All possible technical arrangements have been implemented into the design and the construction to reach this goal. This has been achieved by selecting a quiet tube such as the ECC86/6GM8 as previously stated.

In addition, all the tube heaters are fed by direct current, while the plate voltages are supplied by a dedicated voltage regulator to stabilize its output and remove any possible ripple. You have the option of using external rechargeable batteries for additional noise reduction and overall quieter circuits.

The power supply schematic (**Fig. 4**) shows the hi voltage (+48Vcc) power supply is provided by a simple voltage regulator, while the 6 Vcc is made up of an LM338K, a three-terminal integrated circuit. You can adjust the output voltages on both circuits by means of a dedicated potentiometer.

Battery supplies are activated when mains is off and switch 2 is turned on. Diode 1N4007 on TB3-3 prevents an accident polarity reversal of 48 Vcc external battery to protect all electrolytic caps. Polarity reversal of 6 Vcc battery does not have any adverse effect.

You can load the MC input with a selected resistor to better suit cartridges specifications. A DIP switch allows selection of load values from 12Ω through 1K in 15 steps.

V1 output feeds a passive RIAA network made up of R1, R2, C1, C2. The signal is then applied to V3, V4 buffer stage for an additional 20dB amplification.

V5 functions as cathode follower to provide the lowest possible output impedance; it has no gain. DIYers may elect to eliminate this stage if the connection to a line preamp is kept short (less than a meter).

My implementation uses a high grade toroidal power transformer. The preamplifier is exceptionally quiet when fed with batteries; when powered through an AC outlet, it introduces a certain amount of hum; this, however, in a normal listening environment, does not deteriorate sound quality.

Last, great care must be taken on the grounding scheme by separating signal ground from the power ground wires.

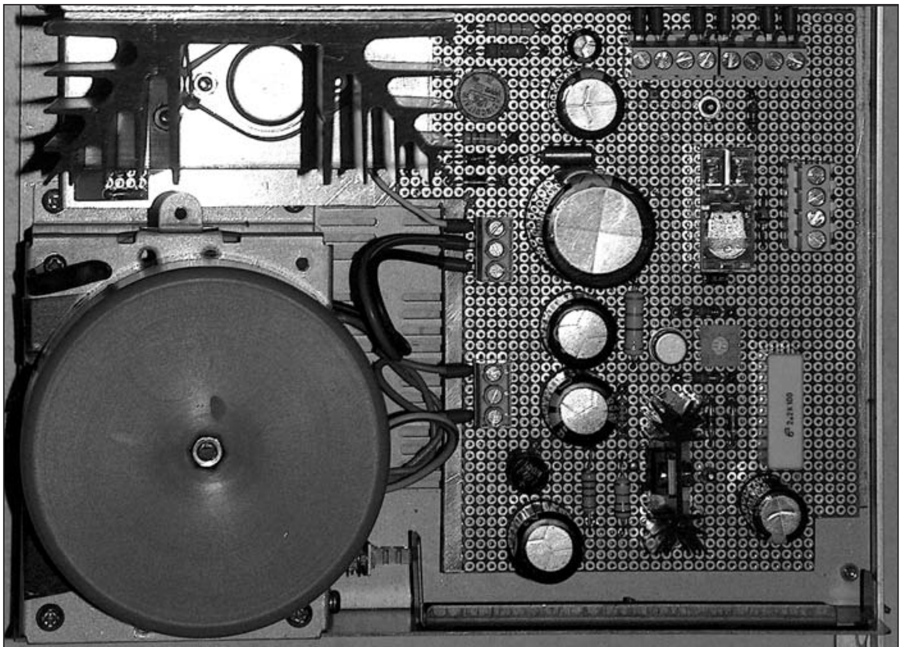

PHOTO 3: Power supply top view. Note the LM338K heatsink and the toroidal transformer positioning.

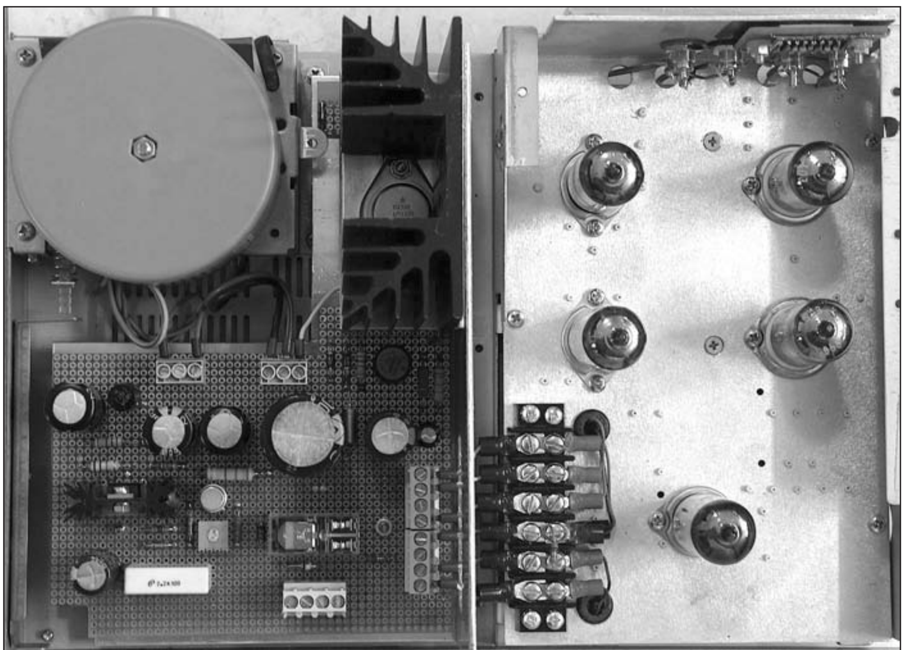

PHOTO 4: Top view of the preamplifier. Box measures 200mm × 280mm × 110mm (D × L × H).

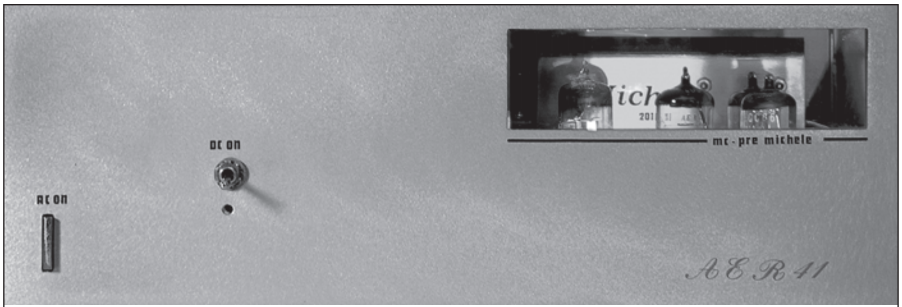

PHOTO 5: Front view. Two bright blue LEDs illuminate the tubes to provide a pleasant effect when seen through the small front window.

PHOTO 6: Back view of preamplifier. On the left input/output RCA connectors and on the right a four-plug receptacle for external batteries. Please note the recycled Fisher cassette deck box.

The detailed schematic shows how this must be done.

Finally, two blue LEDs positioned in front of the tubes produce a nice visual effect when looking through the small front window **(Photo 5).**

CONSTRUCTION

I have reused an old Fisher tape cassette deck as a container and built a small chassis to host all preamp components. All wiring has been made in a point-to-point mode as seen in **Photo 2.** I took great care to keep high impedance paths as short as possible to avoid the use of screened wires. Signal and power grounding wires were kept separately and joined together at the RCA input socket point. Similar grounding arrangements are required in the power supply as properly shown in the schematic.

Batteries, because of their size, are kept outside the box; they feed the preamp through four plug connectors mounted on the back panel **(Photo 6).** Cartridge load can be selected by positioning a DIP switch on the required value. DIYers who own an MC cartridge may elect to install solely the load required, avoiding installation of a DIP switch and related small board **(Photo 7).**

Finally, it is possible to eliminate the

TABLE 1—PREAMPLIFIER, SINGLE CHANNEL. PARTS IN GRAY FOR TWO CHANNELS.

COMPONENT	VALUE	RATING	NOTE
R1	470Ω	1/4W 5%	Metal
R2	100Ω	1/4W 5%	Metal
R3	33Ω	1/4W 5%	Metal
R4	15Ω	1/4W 5%	Metal
R5, 23	1000Ω	1/4W 5%	Metal
R6	4.7kΩ	1/4W 5%	Metal
R7	11kΩ	1/4W 5%	Metal
R8	10Ω	1/4W 5%	Metal
R9, 11, 17, 20, 26	220Ω	1/4W 5%	Metal
R10	1.8kΩ	1/4W 5%	Metal
R12, 19	470kΩ	1/4W 5%	Metal
R13	82kΩ	1/4W 1%	Metal
R14	10.5kΩ	1/4W 1%	Metal
R15	1MΩ	1/4W 5%	Metal
R16, 18	3.9kΩ	1/2W 5%	Metal
R21	6.8kΩ	1/4W 5%	Metal
R22	39kΩ	1/2W 5%	Metal
R24, 25 (see text)	100kΩ	1/4W 5%	Metal
P1	20kΩ	1/4W 20%	Cermet
C1, 7, 10	22μF	100V	Electrolytic
C2, 5, 8, 9 (see text)	1μF	250V 20%	Polypropylene
C3	10nF	250V 1%	Polypropylene
C4	28.4nF	250V 1%	Polypropylene
C6	220μF	25V	Electrolytic
TR1	BC 149		2N3904
L1	6.8μH		RF coil
V1, V2, V3, V4, V5	ECC86		6GM8
DIP Switch	Eight ways		

TABLE 2—POWER SUPPLY PARTS LIST.

COMPONENT	VALUE	RATING	NOTE
R1, 11 (fuse)	0Ω	1/4W 5%	Metal
R2, 5	150Ω	1W 5%	Metal
R3	1.8kΩ	1/2W 5%	Metal
R4	1kΩ	1/2W 5%	Metal
R6	1.2kΩ	1/2W 5%	Metal
R7	3.3kΩ	1/4W 5%	Metal
R8, 9	22kΩ	1/2W 5%	Metal
R10	680Ω	1W 5%	Metal
P1	680Ω	1/2W	Cermet
P2	20kΩ	1/2W	Cermet
D1	W04M	400V 1A	Or equivalent
D2, 3, 4, 5	1N4007	1A 1000V	Or equivalent
D6	B80C5000	80V 5A	Or equivalent
D7	LED		Red
D8	LED		Green
D9, 10	LED		Blue
Z1	24V	1W	Zener
TR1	TIP 102	Darlington	BDX53C
TR2	BFX40	PNP	BSS68
IC1	LM338K	Adj. Regulator	Or equivalent
C1	6800μF	35V	Electrolytic
C2	1000μF	35V	Electrolytic
C3	1.0μF	100V	Polypropylene
C4	100μF	25V	Electrolytic
C5, 6, 7	100μF	100V	Electrolytic
RY1	12V	3A contacts	
T1 Main Transformer	Toroidal	30VA	See schematic
Terminal boards			See schematic
Switches, fuses, etc.			See schematic

V5 cathode follower if connection to a line preamplifier is relatively short. To do this remove V5, R21,R22, R23, C7, and install C9, R5 (by reusing R24 and C8).

ADJUSTMENTS

You should check out the power supplies before making any connection to the preamplifier sections to make sure all voltages are properly set. It would be a good idea to test them by loading each output as follows:

–6Vcc heater PS with 4ΩΩ 10W resistor (1.5A), then adjust P1 680Ω trimmer for 6.3V output.

–48 Vcc with 10kΩ 1W resistor, by adjusting P2 20kΩ potentiometer for proper voltage.

–TR1 bias adjustments (P1 on the preamplifier) must be carried out as follows: with no input signal, measure TR1 collector DC voltage and adjust P1 to read 1.3 –1.35V.

Certainly, the use of batteries eliminates completely any hum from the preamplifier. However, with an audio-grade transformer, and *great care* in the layout and grounding system, you can reach an acceptable level of hum which in practice disappears (masked) when a record is played at any volume level. Have fun. aX

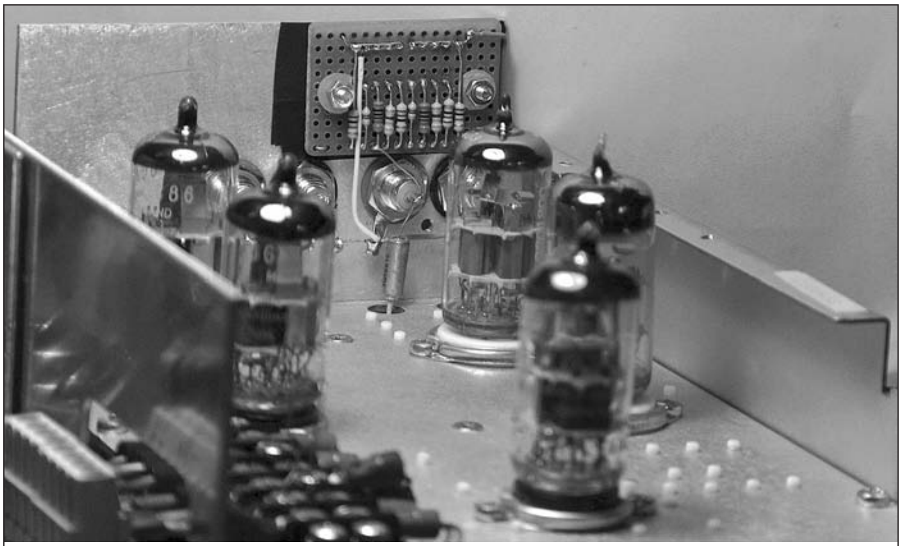

PHOTO 7: Detailed view of DIP switch board. Its position, close to the input sockets, prevents hum pickup.

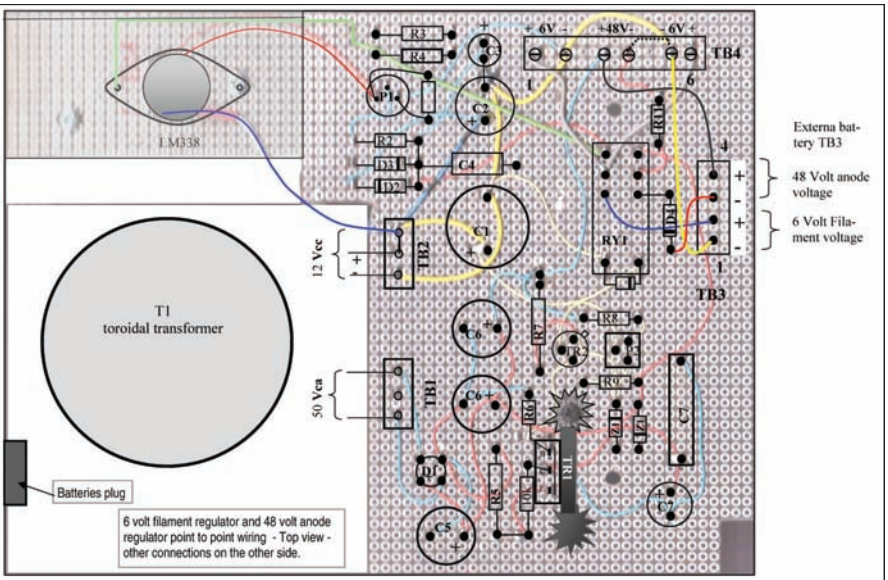

FIGURE 5: Power supply layout. For a more professional and reliable solution, DIYers capable of doing it can develop a PC board to host all power supply components instead of a "multi-holes" board as I did.

By Bill Reeve

Analog Notch Filter

Digital potentiometers enable an adjustable analog notch filter.

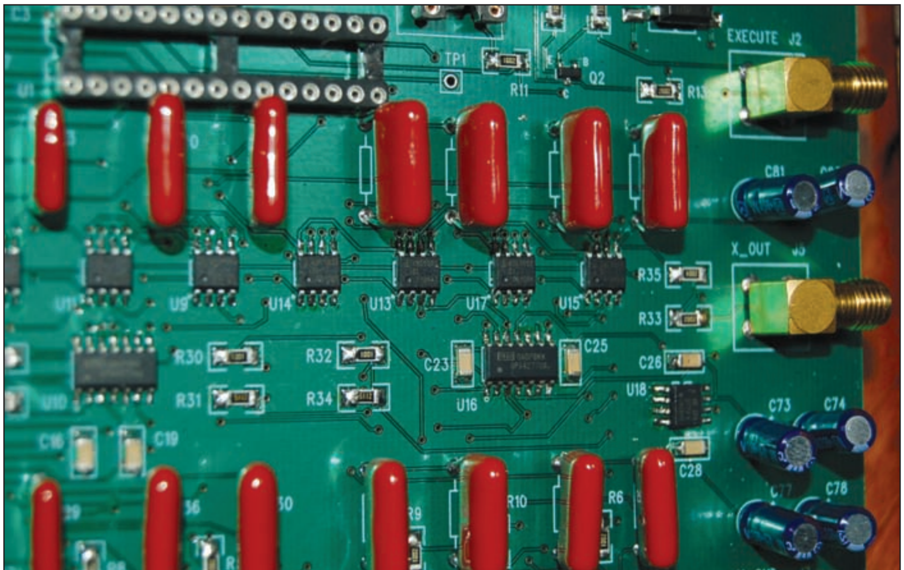

PHOTO 1: Notch circuit on PCB. Red parts are polypropylene capacitors; 8-pin SOIC ICs are digital potentiometers; and 14-pin ICs are quad op amps.

Combining modern digital potentiometers with a classic analog circuit enables a simple processor-controlled adjustable notch (or, alternately, a bandpass) filter that maintains a constant shape (center frequency gain and quality factor "Q") over an adjustable frequency span.

Audio feedback suppression, open-loop harmonic noise elimination, and closed-loop controller performance enhancement are a few applications for this adjustable narrowband filter. Specifically, controllers may need gain peaks below their crossover frequency to suppress harmonic disturbances or mechanical bending modes, or closed-loop performance may benefit from gain notches above the controller's crossover frequency to avoid residual controller gain amplifying higher frequency harmonic signals. This circuit's advantage in these applications is that its out-of-band gain and phase distortion are small, so phase loss does not threaten controller stability.

CIRCUIT DESCRIPTION

Figure 1 is a simplified schematic. The circuit is composed of gain α and a filter op amp. The gain α is implemented with either a non-inverting op amp in notch applications, or by using a voltage divider to obtain a bandpass.

The circuit (**Photo 1**) is a slight adaptation of the classic Delyiannis-Friend filter. It applies gain at a selectable frequency while minimizing gain and phase variation at other (out-of-band) frequencies. The filter op amp's non-inverting terminal is referenced to the input signal to provide unity (0dB) out-of-band gain, so the filter can be applied in series with traditional proportional, integral, and derivative (PID) type controllers.

The filter contains two poles and two zeros, all at the filter's center frequency, and the difference in damping between these poles and zeros produces gain at that frequency. The fact that the poles and zeros are all at the same frequency minimizes gain and phase variation away from the filter's center frequency.

A unique advantage of this formulation is that by setting C1 = C2 and keeping the ratio of R1 and R2 constant (which can be accomplished by electronically "ganging" two digitally-controlled potentiometers of different maximum values), the filter's center frequency can be moved in frequency without transfer function distortion.

The filter transfer function [(output voltage)/(input voltage) = Vout/Vin] is:

$$\frac{V_{out}}{V_{in}} = \left[\frac{s^2 + s(2\xi_n\omega_o) + \omega_o^2}{s^2 + s(2\xi_d\omega_o) + \omega_o^2}\right]$$

The filter center frequency (ω_o), in radians per second, is:

$$\omega_o = 1/\sqrt{R1R2C1C2}$$

The numerator's damping ratio (ξ_n), or the zeros' damping, is:

$$\xi_n = [R1(C1 + C2) + R2C1(1-\alpha)]/2\sqrt{R1R2C1C2}$$

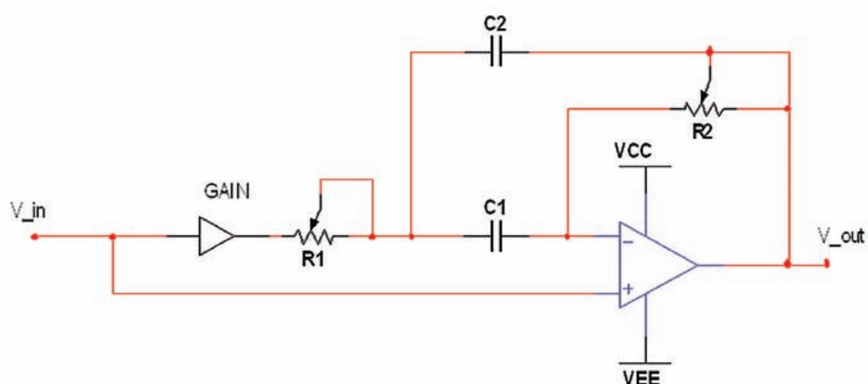

FIGURE 1: Simplified filter schematic.

The denominator's damping ratio (ξd), or the poles' damping, is:

$\xi_d = [R1(C1 + C2)]/2\sqrt{R1R2C1C2}$

Setting the Laplace variable "s" to $j\omega$, and looking at the case where $\omega = \omega_o$ (the filter's center frequency), the transfer function magnitude simplifies to:

$$\left|\frac{V_{out}}{V_{in}}\right| = \left|\frac{\xi_n}{\xi_d}\right| = \left|\frac{Q_p}{Q_z}\right| = \left|1 + \frac{R_2}{2R_1}(1-\alpha)\right|$$

To obtain a reasonable center frequency gain, use a 100K digital pot for R2, and a 1K digital pot for R1. These are commonly available values. By sending them identical digital commands, the ratio R2/R1 is fixed at 100. The desired frequency range is set by choosing the value of the capacitors C1 = C2.

The center frequency gain, and—more important—whether the filter provides a notch or a bandpass, is now controlled by the value of the gain α. **Figure 2** is a plot of the filter's center frequency gain as a function of α over a range of α values between 0.8 and 1.02, assuming the ratio R2/R1 = 100.

APPLICATION

Note that the maximum center frequency gain (theoretically infinity) occurs when α = 1.02. In reality, the notch depth is limited by the filter capacitors' ESR. Use polypropylene or ceramic capacitors for a relatively deep notch. Electrolytic capacitors can perform poorly.

If α is equal to 1, the filter produces no gain (0dB). When α becomes slightly different than 1, gain appears at frequency ω_0. As α moves away from unity, the center frequency gain grows in mag-

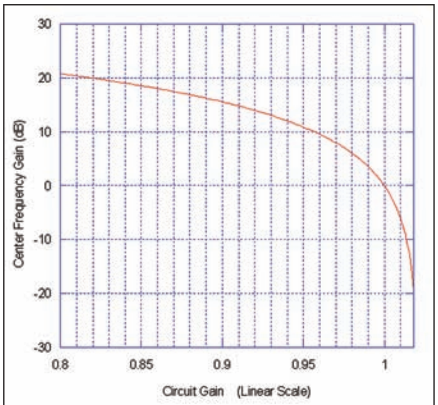

FIGURE 2: Detail of center frequency gain as a function of the circuit gain α, with α slightly below unity.

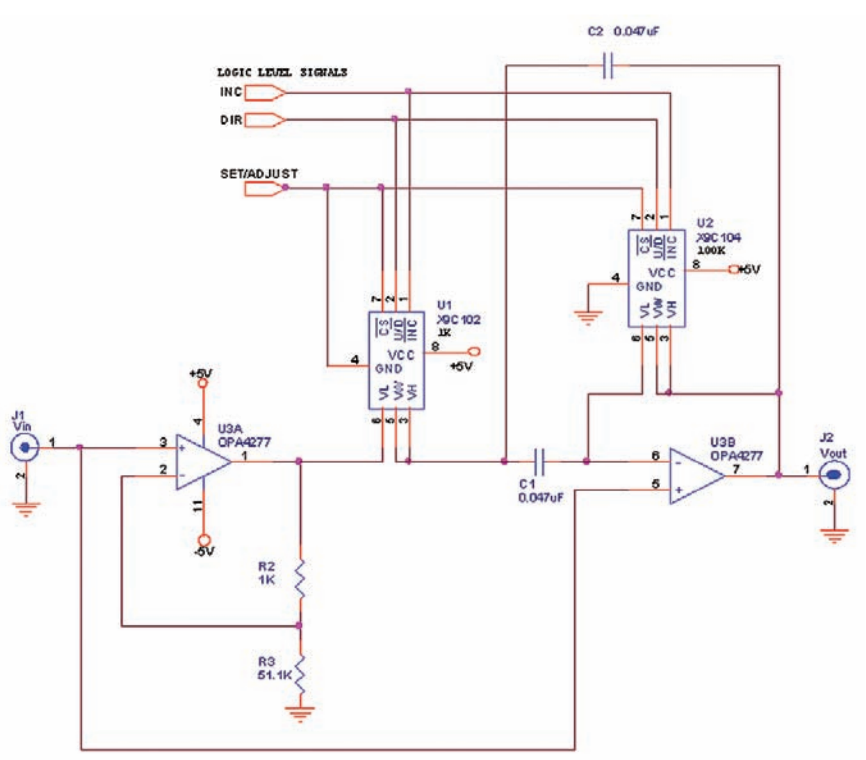

FIGURE 3: Schematic of a digitally adjustable notch filter with good frequency resolution between 340Hz and 1.1kHz (schematic used quad op amp, which could be a dual).

TABLE 1:

Reference designator	Manufacturer's part number	Digi-key part number	Manufacturer	Description
R2*	MFR-25FBF-1K00	1.00KXBK-ND	Yageo	1.00k resistor, 1%, ¼W, metal film, through-hole
R3	MFR-25FBF-51K1	51.1KXBK-ND	Yageo	51.1k resistor, 1%, ¼W metal film, through-hole
C1, C2	QXP2E473KRPT	493-3629-ND	Nichicon	Capacitor, 0.047μF, 10%, high stability, metalized polypropylene, through-hole
U1	X9C102PIZ	X9C102PIZ-ND	Intersil	1k, 100-tap digitally controlled potentiometer, 8-pin DIP
U2	X9C104PIZ	X9C104PIZ-ND	Intersil	100k, 100-tap digitally controlled, 8-pin DIP
U3	OPA4277PA	OPA4277PA-ND	Texas Instruments	Quad op amp, 8-pin DIP

** Author's first resistor value is R2; it could be R1.*

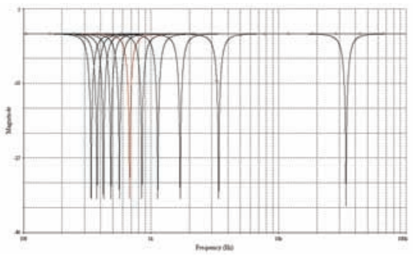

FIGURE 4: SPICE simulation results from the adjustable notch filter shown in Fig. 3. Transfer function magnitude (dB) is plotted vs. frequency showing the filter's characteristic at every tenth setpoint.

nitude and bandwidth. The higher and broader the center frequency gain, the larger the phase variation at the center frequency and the longer the phase takes to return to zero.

Digital potentiometers usually have equally spaced resistance increments, so the filter's resolution is better in the lower region of its adjustable span. Also, as the potentiometer resistances become small (at higher center frequencies), capacitor ESR becomes significant and diminishes the center frequency gain.

Figure 3 is the schematic of an adjustable notch filter with good frequency resolution from about 340Hz to 1.1kHz using non-volatile digital potentiometers. For this circuit, the gain α approximately equals 1.02. Digital signals (SET/ADJUST, DIR, and INC) control the potentiometer values and thus the notch frequency.

Figure 4 shows SPICE simulation results of this filter's transfer function magnitude plotted every tenth potentiometer step. In other words, there are eight settable positions between each of the notches shown in **Fig. 4**. Note the good resolution over the lowest 70% of the frequency span (approximately 340Hz and 1.1kHz).

This architecture is flexible: you can configure the circuit as a notch or pass gain filter by setting the gain α above or below 1; the range of center frequency adjustability can be controlled by capacitor value selection; the filter op amp's non-inverting input can be referenced to ground (rather than Vin) to produce a traditional bandstop or bandpass filter; and the center frequency can be controlled by a processor, programmable logic device, or a manually operated encoder. *aX*

Contributors

Edward T. Dell, Jr. ("Editorial," p. 5), editor/publisher of *audioXpress*, bids farewell after over 40 years of audio-related publishing.

Rudy Godmaire ("Designing Amps for Maximum Linearity," p. 6) works as a senior sales consultant for Bell Canada. Since 1998 DIY audio has been his passion, which he deeply enjoys sharing with other DIYers and especially with friends François and Pierre. Above all, listening to music and attending classical concerts and opera with his wife Elaine remains among the most fulfilling activities he cherishes.

David A. Rich ("Harman-Kardon HD 990" review, p. 13) received his MSEE from Columbia University and his Ph.D. from Polytechnic University of NYU. He specializes in the design of analog and mixed-signal integrated circuits and has taught graduate and undergraduate courses in integrated electronics and electroacoustics. Student work under his guidance, including a novel high-efficiency mixed-signal integrated power amplifier, has won numerous awards. His industrial positions include Technical Manager at Bell Laboratories. His portfolio has spanned the design of audio ICs for Air Force One to RF ICs for wireless cell phones, and his innovations have earned 14 patents. He is a Senior Member of the IEEE and has frequently served as chairperson for technical and panel sessions at IEEE conferences. He has been a member of the AES signal processing technical committee and has been Technical Editor for *Audio Critic*. He is the head of the music committee of the Bethlehem Chamber Music Society.

Jeff Macaulay ("Efficient Class-A Amp," p. 16) first became interested in hi-fi in 1970 when he heard the sound of Jimi Hendrix through a pair of Cambridge R50 transmission line speakers. During a long period of illness, which left him with lots of time, he taught himself to design with transistors. In 1975 he had his first technical article published and soon he was working as a freelance writer. In the mid-1980s he ran a firm providing valve amplifier kits in the UK. More recently he has written for *Electronics World.*

Atto Rinaldo ("The Michele Preamp," p. 20) is a retired 32-year IBM employee who has a keen interest in tubes. He spends some of his time with his grandson Michael, who was the inspiration for this project. Visit his website at www.nuvistor.it.

Bill Reeve ("Adjustable Analog Notch Filter," p. 26) is a Director at Lockheed Martin Corporation in Palo Alto, Calif. He has graduate engineering degrees from the Colorado School of Mines and the University of California, Berkeley as well as a Masters in Electrical Engineering from Santa Clara University.

VINTAGE OCTAL TUBE PREAMP

It's been a good while since we've had an octal-base tube phono preamp, so I thought I would pass along this schematic for the Grommes model 205 PA preamp (**Fig. 1**). The 205PA got its B+ and heater voltage from the Grommes model 117PS power supply. **Figure 1** indicates that 275-325V DC is required for the B+, and, of course, 6.3V for the tube heaters.

Perhaps someone willing to experiment with it will make a great project and share it with the rest of us.

Neal A. Haight
Tracy, CA

HELP WANTED

Does anyone have an idea how to design a zobel for capacitive reactance? I would like to have a means of taming an electrostatic loudspeaker's (ESL) two areas of misbehavior—the high impedance at low frequencies and the low impedance at high frequencies. Any help or advice would be greatly appreciated.

Angel Rivera
Speakerman@excite.com

Readers with information on this topic are encouraged to respond directly to this letter writer at the address provided.—eds.

TURNTABLE TYPES

As a follow-up to my response to Mr. Routh's question about direct-drive versus belt-drive turntables in the June *aX*, there is a review of the Bardo direct-drive turntable in the May 2011 issue of *Stereophile* magazine in which Michael Fremer discusses this question. The problem is many commercial DD turntables are built for DJ use and have high-torque motors and lightweight turntables (TTs) for quick starts and stops. The lightweight TTs do not smooth out the "cogging" steps from the DD motors and thus have moderate to high wow and flutter. A heavier TT and a low-torque motor (fewer poles so less cogging) results in much better performance.

The Bardo (Brinkmann Audio, Achberg, Germany) is highly rated but it costs about $8000 without a tonearm or other accessories. For more details you can read the entire review, which is posted online at www.stereophile.com.

Ron Tipton
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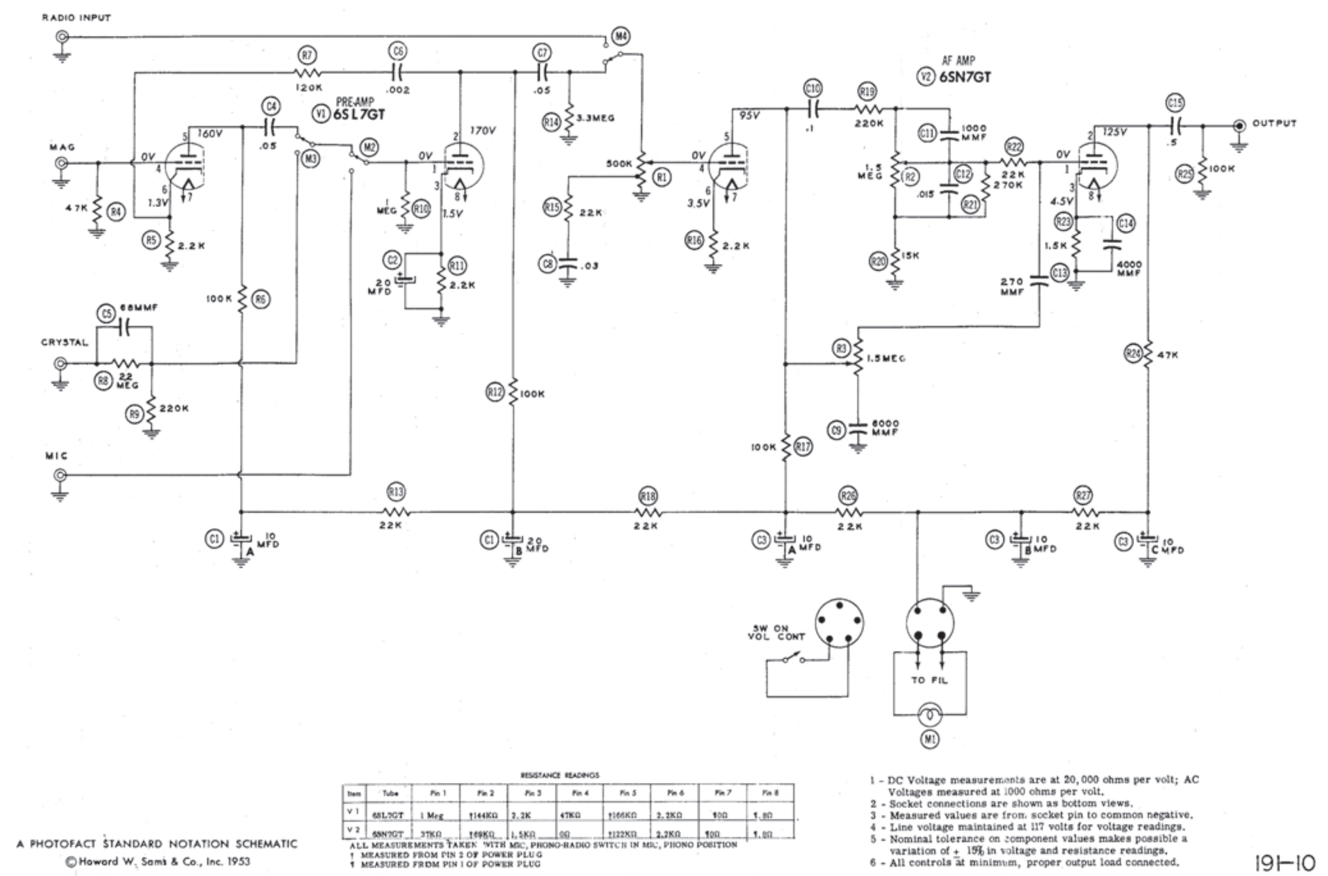

FIGURE 1: The Grommes model 205A PA preamp circuit.

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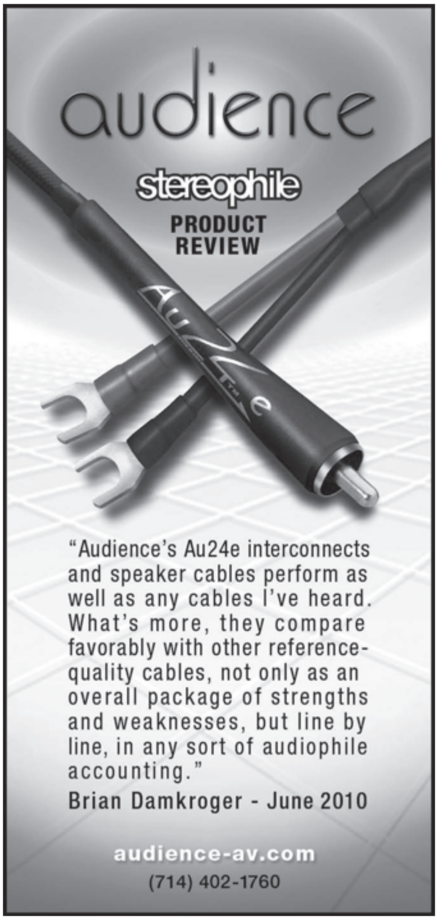

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